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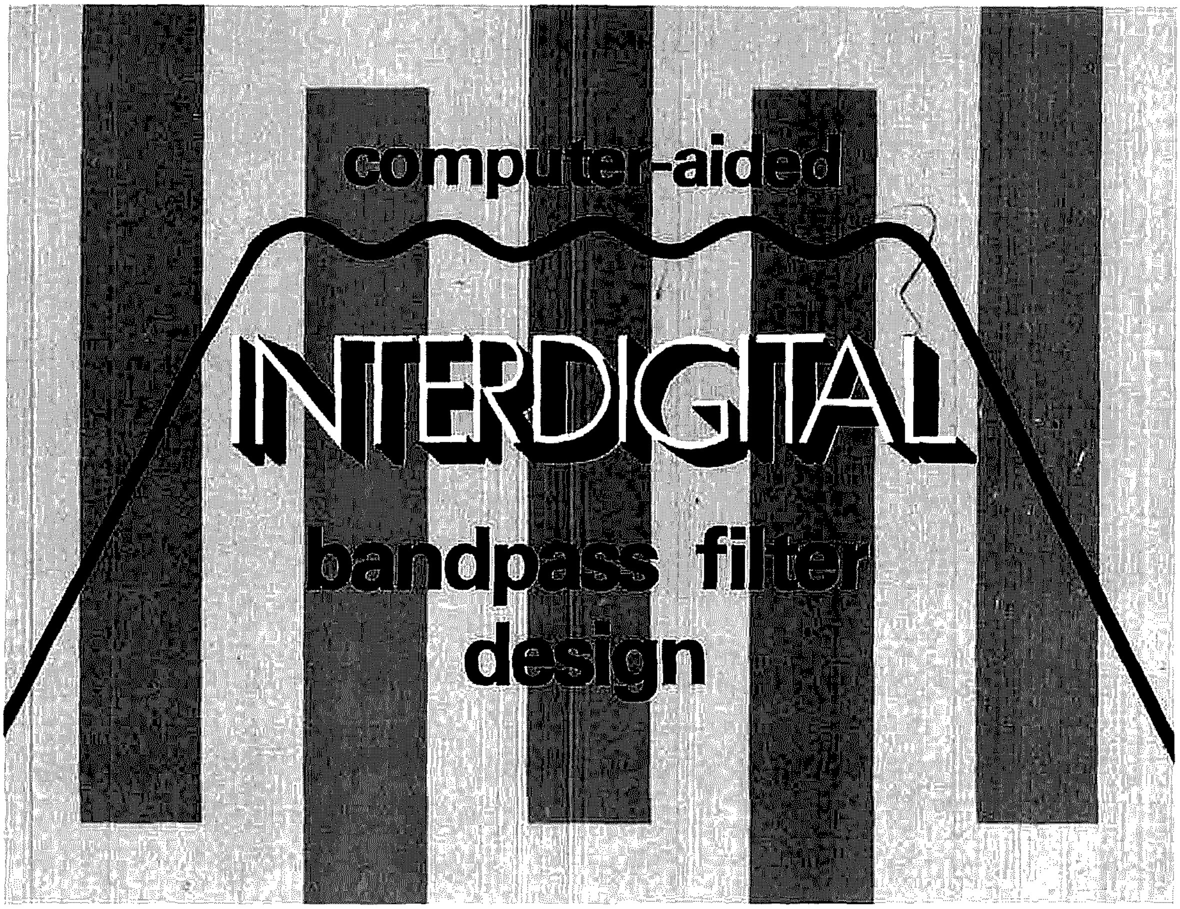
ham radio magazine



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contents

REFLECTIONS

160 meters: spectrum under fire

Once again, Amateur Radio faces a major threat. We recently lost 80 MHz of the 2300-MHz band to aeronautical telemetry. We lost the bottom 10 MHz of 450 MHz within 75 miles of the Canadian border. We lost 25 MHz of the 1200 MHz band to a NAVSAT (navigation satellite) service. We may *yet* lose the 220-MHz band.* Now the 160-meter band is threatened. There's a very real possibility that a proposal to allocate the top half of 160 (1.900 – 2.000 MHz, Amateur exclusive) to the offshore navigation service will be adopted by the FCC.

If you haven't been on 160 during the past few years, you might to surprised by what you would hear now. On any given evening you'll find plenty of stations conducting both CW and SSB ragchews. Later, the DX'ers appear.

Maybe you think 160 is not very important because you never operate there. But consider this: for every *two* stations on 160, there's that much more space available on one of the other bands for you to use. Ten years ago, 160 was considered the AMers' band and there was little activity on it. Not much commercial equipment covered the band. But now, try 160 during any major contest weekend; you'll find plenty of stations on and plenty of DX to be worked.

The FCC's proposal to move the offshore navigation service from 1.6-1.8 MHz to 1.9-2 MHz is predicated on the WARC decision to expand the AM broadcast band up to 1.705 MHz. Industry pundits have suggested that this expansion may end up being more of a boondoggle than a benefit to the consumer and the broadcast industry. There are millions of AM radios around — few cover the entire 1.6-1.750 MHz band!

Some serious questions need to be answered. Can the broadcast industry and the general business community financially support additional AM stations? Competition is tough enough now. Will other users of the spectrum be adversely affected in any way? It's been said that the proposed expansion is the result of some political debt being repaid. Whatever the motive, the proposal seems to be a foolhardy and unnecessary exercise.

Now what about the offshore navigation interests? Serious questions have been raised about this service. Why is it necessary to have an MF radiolocation service when there are other, more precise methods of radiolocation such as LORAN-C and NAVSAT? How about using technology similar to that of the new microwave landing system currently being integrated into airports by the FAA? The bottom line is that offshore navigation has few, if any, credible reasons to be moved into the 1.9-2 MHz slot. If the broadcast band is to be expanded only as far as 1.705 MHz, can anyone explain why, with selective receivers and stable transmitters, the offshore navigation interests cannot be accommodated between 1.705 and 1.8 MHz? It's almost the same size as the allocation proposed, and offshore navigation already operates there. Furthermore, if offshore navigation interests are allowed to get their way, what's to stop other services from claiming Amateur frequencies on the basis of equally flimsy justification? The ultimate result would surely be a major disruption in Amateur Radio as we know it. Don't kid yourself. This could be just the beginning!

If you haven't already filed your comments with the FCC on the proposed 160-meter reallocation, now is the time to do so. Your comments must be received by the FCC before January 24, 1985. Be reasonable and concise. Give solid technical and operational reasons as to why this proposal should not be accepted. Write "Docket 84-874" at the top of each page of your comments, include five additional copies (eleven if you wish each commissioner to have one), and send them to THE SECRETARY, FCC, Washington, D. C. 20554.

It is a unique privilege that we have to be able to be included in the decision-making process proposals such as this. If you don't sit down right now, write up your comments, and send them in, then when the FCC accepts the offshore navigation proposal because of a lack of adequate opposition from the Amateur community, all that can be said is "Quityergripen!"

J. Craig Clark, N1ACH
Assistant Publisher

*See "Reflections," *ham radio*, October, 1984 (page 6) and this month's *Presstop* (page 8).

presstop de W9JUV

COMMERCIAL PRESSURE ON THE 220-MHZ BAND HAS BEEN RELIEVED by the FCC's assignment of the Land Mobile reserve frequencies between 806 and 947 MHz to that and other services November 21. In its far-reaching action, the Commission rejected both Mura's proposal for a new Personal Radio Service for 8 MHz and a commercial proposal to allocate 896-898 and 941-943 MHz to an air-ground telephone service; GE's proposal for a Personal Radio Service in the 900 MHz region was dropped just a few weeks earlier.

12 MHz For Land Mobile And 12 MHz For Cellular Were Proposed in the November 21 meeting, while 6 MHz was reallocated to government and non-government fixed services and frequencies in the 944-947 MHz band were reallocated for broadcast links and relays. As the "220 Grab" was already in trouble from a variety of directions, this action is likely to effectively remove 220 from Land Mobile's sights. However, it's also just as likely to create a whole new set of user pressures on 220 MHz as well as other Amateur bands.

The Failure To Establish Any Form Of A "Personal" Radio Service in the Land Mobile reserve frequencies is regarded as a severe defeat by would-be users and suppliers alike. The concept had been strongly supported by the GMRS* community. The next logical targets for a Personal Radio Service could now be 902-928 MHz, and (repeating history) 220 MHz! The air-to-ground telephone proponents could, on the other hand, decide to go after 420-430 MHz for their new service since that band has already gone to Land Mobile in Canada and is being protected for Canadian benefit along the Canadian border.

Whether Or Not These Fears Materialize, pressure on Amateur frequencies is expected to continue. The recent loss of the Amateur secondary allocation of 2310-2390 MHz to flight telemetry is a case in point. The current stagnation of Amateur growth, coupled with apparent FCC coolness toward the Amateur Service, could spell trouble ahead.

TWO LONG-TERM ARRL DIRECTORS WERE UNSEATED in the League elections. In the New England Division Tom Frenaye, K1KI, soundly beat John Sullivan, W1HHR. In a closer race Linda Ferdinand, N2YL, beat out incumbent George Diehl, W2IHA, in the Hudson Division. Incumbents won in all other races in which they ran; however two new Vice Directors, Rush Drake, W7RM, (Northwestern Division) and Wayne Overbeck, N6NB, (Southwestern Division) were both also elected.

EXTENSIVE AMATEUR OPERATION DURING A 1985 SPACE SHUTTLE mission now seems certain, according to word from NASA. It's to be on mission 51-F, now scheduled for next April, though delays in the Shuttle schedule could push that back to July or even later. Two Amateurs are scheduled for mission 51-F, and it appears likely there'll be operation on a variety of modes and frequencies with some sort of repeater also possible.

ARRL'S LEAVING NEWINGTON AND A MAJOR PUSH FOR AMATEUR GROWTH were just two of the many significant items covered by the League's directors at their fall meeting. As for moving, the Management and Finance Committee was directed to conduct a study of possible sites, costs, timing, and impact on the membership and staff. Though not specified in the board minutes, Washington, D.C. is believed to be what's in mind. A growth of 50,000 new Amateurs a year for the next five years is the goal of a new plan devised by the League's General Manager, for implementation in 1985. (The number of individual U.S. licensed Amateurs has held almost steady at just over 400,000 for the past several years.) He's also been instructed to develop a parallel program to increase ARRL membership by 25,000 in 1985, and by 20% per year thereafter. Other items included an apparent endorsement of simplex autopatches, a membership survey of further phone expansion on 7 MHz, and a Plans and Programs Committee study on the League becoming involved with maintenance of Amateur licensing records and "especially in the administration of special call sign requests...."

ARRL'S REQUEST FOR FCC PREEMPTION OF ANTENNA REGULATIONS, PRB-1, has been drawing a lot more favorable—and effective—Amateur comment. November 9 the FCC extended the comment cutoff date to December 24, and indications are that the flow of comments from Amateurs describing the problems they've had with local restrictions is continuing at a good rate.

What The Outcome Will Be, However, Remains To Be Seen. There have also been some well presented arguments from the other side of the fence, and the issue is one which does have two valid sides. Which side, if either, the Commissioners will take cannot be predicted.

FCC'S PROPOSAL TO REALLOCATE 1900-2000 KHZ to radio direction finding is still open for comment. Just after last month's Presstop went to press, the Commission granted an ARRL request for an extension of the Comment due date to January 24, 1985, with Reply Comments due March 11. An original and five copies sent to the Secretary, FCC, are needed for a formal filing; refer to PR Docket 84-874. (See "Reflections," page 5.)

440-450 MHZ IS FULL IN SOUTHERN CALIFORNIA, according to the Southern California Repeater and Remote Base Association (SCRRBA), with all repeater channels filled. Unless SCRRBA's review of activity turns up some inactive systems, would-be repeater operators will have to go to 23 cm since 2 meters and 220 have long been filled in that area.

*General Mobile Radio Service. Formerly "Class A" CB, GMRS has developed into a sophisticated Personal FM Radio Service with frequencies in the 460-MHz band.

computer-aided interdigital bandpass filter design

For reproducible customized design in the 0.4-5 GHz frequency range

One of the more challenging difficulties facing the Amateur experimenter in UHF and microwave communications is to construct good bandpass filters at these high frequencies. Good filters are especially important in those bands where the Amateur frequency allocation is shared with or is near high power services, such as radar, which can cause ruinous interference. And although a number of articles have appeared in Amateur publications describing specific bandpass filter types and specific construction techniques, none has provided a simple means of designing filters for different requirements.

This article describes a flexible computer program that makes it possible for Amateurs to design and build their own bandpass filters, custom-tailored to their specific needs. The program performs the design tasks for interdigital filters, a common bandpass type. The article also shows, step by step, how to proceed from the computer printout to the building and testing of a working filter. In addition, two different examples are used to illustrate two different construction methods.

filter design

The design of narrowband bandpass filters — that is, of filters that have passbands of approximately 10 percent of the center frequency or less — is often based on the well-known work of Matthaei, Young, and Jones,[1] which provides a wealth of analytical and practical design methods for many types of RF and microwave filters. Among the most common types are filters that use comb or interdigitated coupled resonators. These two types are mechanically similar, but this article discusses only the interdigital form.

Interdigitated bandpass filters have been widely used in the microwave electronics industry for many years. These filters are commonly used because they provide reasonably good passband characteristics, moderate loss, and fairly high attenuation in the stopbands. Furthermore, they are simple to build and tune.

A typical interdigitated bandpass filter, such as shown in **fig. 1**, consists of a number of resonator elements or rods, each approximately a quarter wavelength long at the center frequency of the filter, which are electrically coupled together between two conducting ground planes. Each rod is shorted to ground at one end and open-circuited at the other end. The rods alternate, with one rod's shorted end opposing the next rod's open end. It is this alternating structure, which looks somewhat like interlaced fingers, that gives the interdigital filter its name. At the ends of the filter some form of impedance matching, either a transmission line transformer or a tap on the end rods, is used to couple energy into and out of the filter.

Interdigitated filters are most useful in the low microwave frequency range of about 0.5 to 5 GHz. In this region, lumped element filters are difficult to build, and waveguide filters are mechanically large. The inter-

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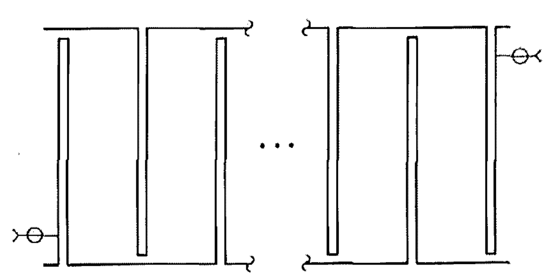

fig. 1. Typical interdigital bandpass filter structure. This filter consists of a number of rods, each open circuited at one end and grounded at the other, which are electrically coupled together between ground planes.

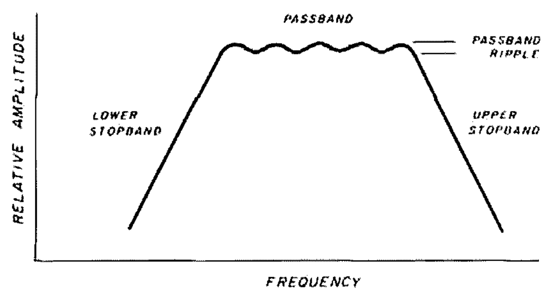

fig. 2. A generalized bandpass filter response. The ripple in the passband is exaggerated for clarity.

digital filter handily fills this gap between low frequency "coil and capacitor" filters, and microwave waveguide "plumbing." Thus, interdigital filters are of interest for frequencies up to at least several GHz, and down at least as low as the 420 MHz ham band.

The traditional design of interdigitated filters described by Matthaei, Young, and Jones calls for both the spacing and sizes of the rectangular resonant elements to be variables. While there is no problem with this theoretically, in practice it is often simpler to use round rods of equal diameter in place of rectangular ones of various sizes. In the 1960's, Dishal[2] described a method of designing narrow bandwidth filters using equal diameter round rods. His method provides a simple and accurate design guide which results in bandpass filters of straightforward mechanical construction and good electrical performance.

In addition to using uniform round rods in place of rectangular elements of various sizes, Dishal questioned the common practice of using additional elements, one at each end of the filter structure, whose only purpose was to match the filter to the desired input and output connections when a simple tap on the first and last element would serve as well. Taking it

defining terms

A *bandpass filter* is a device that permits the passage of signals within a certain range of frequencies only, but blocks out signals from above or below that range. The range of frequencies where signals can pass through the filter with low loss is called the *passband*. The frequencies outside the passband are variously referred to as the *rejection bands, stop bands* or *"skirts"* of the filter. (**Figure 2** shows a general filter response.)

An ideal bandpass filter would have zero loss in the passband and infinite rejection of undesired signals in the stopbands. In real life, however, the situation is not as clear-cut: the passband of a real filter has a certain finite loss, even though it may be quite low. Likewise, a real filter's stopbands do not absolutely reject signals, but merely reduce their amplitudes. The attenuation depends mainly upon the separation, in frequency, of the undesired signal from the passband.

The loss in the passband may be nearly constant, or flat, as in a Butterworth design, or it may vary with regular undulations or "ripple" in the attenuation curve. In a Chebyshev type filter, this passband ripple is related to an increase in stopband rejection. For our purposes, it is enough to consider a Butterworth filter as a variation of the Chebyshev design, but with zero passband ripple. It is important to realize that we can obtain better stopband rejection if we're willing to trade off passband flatness and accept higher ripple and VSWR.

Since the transition from passband to stopband in a real (rather than ideal) filter is not abrupt, we must define its location. For a Chebyshev filter, the "edges" of the passband are defined as those points where the attenuation exceeds the maximum ripple amplitude. Thus the passband of a Chebyshev filter is often known as the *ripple bandwidth*. The bandwidth of a Butterworth design, which has no ripple, is traditionally defined at the 3 dB points. This measure is called the *3 dB bandwidth*.

There are many good texts and articles dealing with filter topics, and the interested reader should refer to them if a better background is desired. A few of these sources are listed in the references. However, the terms and concepts introduced above are sufficient for a basic understanding of the design and construction of the bandpass filters we describe.

table 1. Interactive program calculates expected electrical performance and computes mechanical dimensions of filter.

```
10 REM ***********************************************
20 REM   DESIGNS INTERDIGITAL BPFs
30 REM ***********************************************
40  DEF FNRJ(TA,B,C,D)=(B*C-TA*D)/(C*C+D*D)
50 REM using equal diameter rods
60 REM g values based on ripple bw . q/coup on 3-bd
70 DIM G(200) , C(200) , RK(200) , AK(200) , FR(40) , ALOSS(40)
80  DIM A(200), B(200)
90 PI=3.14159265#
100 INPUT"# OF ELEMENT $ P-P RIPPLE IN PASSBAND (DB)";N,RIP
110 REM
120 INPUT"INPUT FILTER CENTER FREQ.(GHZ),BW(MHZ)&LOAD IMPEDENCE
     ZO";FZGC,BWMC,R
130 REM
140 PRINT"INPUT GROUND PLANE SPACING , ROD DIAMETER"
150 INPUT"& DISTANCE TO CENTER OF FIRST AND LAST ROD";H,D,E
160 REM
170 REM
180 INPUT"NO. OF FREQ. REJECTION PTS AND STEP SIZE (MHZ)";NFR,STP
190 FOR IP=-NFR/2 TO NFR/2
200      CONTER=CONTER+1
210      FR(CONTER)=FZGC+(STP*.001*IP)
220 NEXT IP
230 IDAT=1
240 GOTO 250
250 F1=FZGC-.0005*BWMC
260 F2=FZGC+.0005*BWMC
270 IF RIP>0 THEN GOTO 330
280 BW3CC=F2-F1
290 BWRGC=0
300 BW3=1
310 GOSUB 1960
320 GOTO 390
330 B=1/SQR(10^(.1*RIP)-1)
340 CA=LOG(B+SQR(B*B-1))/(N)
350 BW3=(EXP(CA)+EXP(-CA))/2
360 GOSUB 1740
370 BWRGC=F2-F1
380 BW3GC=BWRGC*BW3
390 REM
400 W=2*(F2-F1)/(F2+F1)
410 QF=FZGC/BW3GC
420 NFM=N-1
430 QWVL=11.8028/(4*FZGC)
440 FOR K=1 TO NFM
450      AK(K)=1/(BW3*SQR(G(K)*G(K+1)))
460      RK(K)=AK(K)/QF
470 NEXT K
480 AKO=G(1)*BW3
490 AK(N)=AKO
500 AK(N+1)=0
510 QS=G(1)*BW3*QF
520 CANH=(EXP(2*PI*E/H)-1)/(EXP(2*PI*E/H)+1)
530 ZM=59.9585*LOG(4*H/(PI*D))
540 ZE=59.9585*LOG(CANH*H*4/(PI*D))
550 RKM=RK(1)*SQR(ZM/ZE)
560 Z=PI*D/(2*H)
570 COTH=(EXP(Z)+1)/(EXP(Z)-1)
580 Y=PI*RKM/4
590 T=COTH^Y
600 C(1)=(H/PI)*LOG((T+1)/(T-1))
610 MFL=N-2
620 REM IF N-3<0 THEN AG=1 ELSE IF N-3=0 THEN AG=2 ELSE AG=3
630 ON (2+SGN(N)*1) GOTO 690 , 690 , 640
640 FOR K=2 TO MFL
650      Y=PI*RK(K)/4
660      T=COTH^Y
670      C(K)=(H/PI)*LOG((T+1)/(T-1))
680 NEXT K
690 C(N-1)=C(1)
700 X=SQR(PI*R/(4*ZE*QS))
710 AQ=2*QWVL*ATN(X/SQR(1-X*X))/PI
720 QU=2200*H*SQR(FZGC)
730 SUMG=0
740 FOR J1=1 TO N
750      SUMG=SUMG+G(J1)
760 NEXT J1
770 BLOSS=4.34*FZGC*SUMG/(QU*(F2-F1))
780 DELAY=SUMG/(2*PI*(F2-F1))
790 IF RIP > 0 THEN GOTO 820
800 PRINT"DESIGN DATA FOR ";N;" POLE INTERDIGITAL FILTER.
     BUTTERWORTH RESPONSE"
810 GOTO 830
820 PRINT "DESIGN DATA FOR";N;"POLE INTERDIGITAL FILTER .BAND
     PASS RIPPLE";RIP;"DB"
830 PRINT"CENTER FREQ. ";FZGC ;"GHZ"
840 PRINT"CUTOFF FREQ. ";F1;" (GHZ) AND ";F2;"  GHZ"
850 PRINT"RIPPLE BW.  ";BWRGC;"GHZ"
860 PRINT"3 DB BW.    ";BW3GC;"GHZ"
870 PRINT"FRACTIONAL BW.";W
880 PRINT"FILTER Q  ";QF
890 PRINT"EST QU    ";QU
900 PRINT"LOSS BASED ON THIS QU  ";BLOSS;"  DB"
910 PRINT"DELAY AT BAND CENTER ";DELAY;"NANOSECONDS"
920 FOR JK=1 TO NFR
930     IF JK=1 THEN PRINT "FREQUENCY REJECTION INFORMATION "
940     NFN=ABS(2*(FR(JK)-FZGC)/(W*FZGC))
950     IF RIP >0 THEN GOTO 980
960     ALOSS(JK)=10*LOG(1+NFN^(2*N))/LOG(10)
970     GOTO 1040
980     IF NFN<1 THEN NFN=1
990     ANG=N*LOG(NFN+SQR(NFN*NFN-1))
1000     YAK=.5*(EXP(ANG)+EXP(-ANG))
1010     ALOSS(JK)=10*LOG(1+(10^(.1*RIP)-1)*YAK*YAK)/LOG(10)
1020     IF ALOSS(JK)>65 THEN ALOSS = 65 ELSE ALOSS = ALOSS(JK)
1030     FR=INT(FR(JK)*10000)/10000 : ALOS=INT(ALOSS(JK))
1040     PRINT TAB(INT(ALOSS))"*";TAB(66)FR;TAB(73)ALOS
1050 NEXT JK
1060 WO=2*PI*FZGC*1E+09
1070 F=D/H
1080  CF=(-.0000422+.0857397*F+.0067853*F*F-9.092165E-
     02*F^3+.169088*F^4)*PI*H*2.54
1090 REM
1100 WW=WO*1E-12
1110 B2=PI*AQ/(2*QWVL)
1120 GG=1/R
1130 BB=-COS(B2)/(ZE*SIN(B2))
1140 EL1=.8*QWVL
1150 ANG=EL1*PI/(2*QWVL)
1160 B1=ANG-B2
1170 YL=-COS(ANG)/(ZM*SIN(ANG))
1180 CP=WW*(CF+.17655*D*D/(QWVL-EL1))
1190 Y1=CP+YL
1200 EL2=.87*QWVL
1210 ANG=EL2*PI/(2*QWVL)
1220 B4=ANG-B2
1230 YL=-COS(ANG)/(ZM*SIN(ANG))
1240 CD=WW*(CF+.17655*D*D/(QWVL-EL2))
1250 Y2=CD+YL
1260 EL3=.95*QWVL
1270 ANG=EL3*PI/(2*QWVL)
1280 B5=ANG-B2
1290 YL=-COS(ANG)/(ZM*SIN(ANG))
1300 CQ=WW*(CF+.17655*D*D/(QWVL-EL3))
1310 Y3=CQ+YL
1320  ELEM=Y3*Y2*EL1/((Y1-Y2)*(Y1-Y3))+Y1*Y3*EL2/((Y2-Y1)*(Y2-
     Y3))+Y1*Y2*EL3/((Y3-Y1)*(Y3-Y2))
1330 TANN=SIN(B1)/COS(B1)
1340 YL=FNRJ(GG,BB+TANN/ZE,1-ZE*BB*TANN,ZE*GG*TANN)
1350 Y1=CP+YL
1360 TANN=SIN(B4)/COS(B4)
1370 YL=FNRJ(GG,BB+TANN/ZE,1-ZE*BB*TANN,ZE*GG*TANN)
1380 Y2=CD+YL
1390 TANN=SIN(B5)/COS(B5)
1400 YL=FNRJ(GG,BB+TANN/ZE,1-ZE*BB*TANN,ZE*GG*TANN)
1410 Y3=CQ+YL
1420  ELEQ=Y3*Y2*EL1/((Y1-Y2)*(Y1-Y3))+Y1*Y3*EL2/((Y2-Y1)*(Y2-
     Y3))+Y1*Y2*EL3/((Y3-Y1)*(Y3-Y2))
1430 REM
1440 PRINT"QUARTER WAVELENGTH =";QWVL ;"INCHES"
1450 PRINT"THE LENGTH OF INTERIOR ELEMENTS =";ELEM;" INCHES"
1460 PRINT"LENGTH OF END ELEMENTS =";ELEQ;" INCHES"
1470 PRINT"GROUND-PLANE SPACE =";H," INCHES "
1480 PRINT"ROD DIAMETER   =";D;" INCHES"
1490 PRINT"END PLATES";E;" INCHES FROM C/L OF END ROD "
1500 PRINT"TAP EXTERNAL LINES UP ";AQ;" INCHES FROM SHORTED END "
1510 PRINT"LINE IMPEDANCES: END ROD";ZE;" ,OTHER ";ZM;" , EXT.
     LINES ";R;"OHM"
1520 PRINT"DIMENSIONS"
1530 PRINT"EL. NO.     END TO C     C TO C    G(K)       Q/COUP"
1540 DOM=E
1550 GOO=1
1560 PRINT "0";TAB(41)GOO;TAB(55)AKO
1570 PRINT "1";TAB(16)E;TAB(41)G(1);TAB(55)AK(1)
1580 FOR K=1 TO NFM
1590     L=K+1
1600     PRINT TAB(28)C(K)
1610     DOM=DOM+C(K)
1620     PRINT L;TAB(16)DOM;TAB(41)G(L);TAB(55)AK(L)
1630 NEXT K
1640 LQ=N+1
1650 PRINT LQ;TAB(41)G(LQ)
1660 DOM=DOM+E
1670 PRINT TAB(16)DOM
1680 IF IDAT =1 THEN GOTO 2070
1690 REM
1700 REM
1710 REM DEFINE FUNCTION
1720 DEF FNRJ(TA,B,C,D)=(B*C-TA*D)/(C*C+D*D)
1730 END
1740 REM SUB CHEB
1750 REM
1760 C=2*RIP/17.37
1770 BETA=LOG((EXP(C)+1)/(EXP(C)-1))
1780 GAMMA=.5*(EXP(BETA/(2*N))-EXP(-BETA/(2*N)))
1790 FOR K=1 TO N
1800     A(K)=SIN(.5*(2*K-1)*PI/N)
1810     B(K)=GAMMA^2+SIN(K*PI/N)^2
1820 NEXT K
1830 G(1)=2*A(1)/GAMMA
1840 FOR K=2 TO N
1850     G(K)=4*A(K-1)*A(K)/(B(K-1)*G(K-1))
1860 NEXT K
1870 NN=N/2
1880 NNN=(N+1)/2
1890 REM IF NNN-NN<0 THEN AG1=1 ELSE IF NNN-NN=0 THEN AG1=2 ELSE AG1=3
1900 ON (2+SGN(NNN-NN)*1) GOTO 1910,1910,1930
1910 G(N+1)=((EXP(BETA/2)+1)/(EXP(BETA/2)-1))^2
1920 RETURN
1930 G(N+1)=1
1940 RETURN
1950 END
1960 REM SUB FOR BUTT
1970 REM
1980 REM
1990 REM
2000 REM
2010 POV2=1.57079633#
2020 FOR K=1 TO N
2030     G(K)=2*SIN(POV2*(2*K-1)/N)
2040 NEXT K
2050 G(N+1)=1
2060 RETURN
2070 END
```

number of elements	ripple dB	bandwidth MHz	loss at 440 dB	loss at 445 dB
2	0.25	2	1.2	21.6
2	0.25	3	0.8	14.4
3	0.25	2	2.5	41.4
3	0.25	3	1.7	30.5
3	0.25	4	1.3	22.6
4	0.50	1	7.9	88.8
4	0.50	3	2.7	50.0
4	0.25	3	2.4	46.8
6	0.25	3	4.2	79.4
6	0.25	4	3.2	63.4

fig. 3. Program tests various configurations for the 440-MHz bandpass filter. For comparison, all are for 0.38 inch diameter rods between groundplanes 1 inch apart, and all have a center frequency of 440 MHz.

a step further, he showed how to determine these tap locations. A computer program, described below, is developed around this design approach.

program description

The BASIC listing is given in **table 1**. This program follows in loose form a program originally written in Fortran IV by Rook and Taylor.[3] We translated into BASIC, modified it for use on a personal computer, and added additional plotting output. The program uses an interactive approach to request design information from the user, and then calculates the expected electrical performance and computes the mechanical dimensions of the filter. It is written in BASIC in its IBM PC™ version, but it is structured so that conversion into other versions of BASIC for different computers should not be difficult.

The first portion of the program sets up the required variable dimensions and types. Next, an interactive question sequence collects the input data for the design. Once the required data are available to the program, it computes the expected electrical performance and gives details of the filter's mechanical construction. Finally, the program prints a graph of the passband and rejection skirts of the filter.

Two different examples of filters for two different ham bands are given. Each example gives general mechanical details of construction techniques which have been proven to give good results. The examples are just that, examples, and serve as guides to help the reader design and build filters which are optimized for his own particular application. The explanations of these techniques give enough information so that the examples themselves can be duplicated without too much difficulty.

440 MHz bandpass filter

The first example is a front-end filter for a 440 MHz FM receiver in a linear translator system. A linear translator, like a repeater, retransmits what it receives on a frequency 5 MHz (the 440 MHz spacing) away from the input channel. In order to prevent receiver overload or excessive intermodulation distortion in the presence of the strong transmitted signal, the rejection of the near-by repeater transmitter at 445 MHz must be approximately 50 dB. At the same time, the filter's passband loss should be moderate, less than about 3 dB, or the receiver sensitivity will be degraded excessively. These two considerations dictate the choice of filter type and design.

An interdigital bandpass filter turns out to be a good choice for this application. It can be simply and inexpensively built, and has relatively low passband losses together with good out-of-band rejection. And while it is true that a cavity diplexer filter of the sort often seen in amateur repeaters can give still lower losses and greater rejection, its good electrical performance can be accomplished only at the expense of increased mechanical complexity, larger size, and higher cost. Here, the required performance does not absolutely dictate the use of a multiple cavity diplexer, so we should definitely consider using the much simpler interdigital filter. If still less skirt rejection and somewhat greater loss can be tolerated, a helical resonator filter might be a good choice, for it would provide less rejection and probably would have higher passband losses, but it would also be much smaller mechanically than an equivalent interdigital bandpass filter. This size reduction is possible because the inductances in a helical filter are coils and because the inter-element couplings are not dependent only on the physical spacing of elements.

The first step in the choice of filter parameters is to determine the required passband loss limit, passband ripple, and out-of-band rejection that can be tolerated. An interactive computer-aided design program makes these tradeoffs considerably simpler to evaluate. The designer begins by entering a first estimate, or guess, of the approximate number of elements and passband ripple. From these inputs, the computer program quickly determines the approximate loss and plots the pass and reject bands. In the course of a few minutes' work, several different configurations can be tried out. From these it is easy to select the optimum design.

As an example of this interactive optimization, fig. 3 lists a number of different 440 MHz filter configurations tested by the program. These designs differ mainly in the number of elements and in their passband widths. Filters with from 2 to 6 elements and ripple bandwidths of from 1 to 4 MHz are compared. In each case, the loss at the center frequency, 440 MHz, and at the transmitter frequency to be rejected, 445 MHz, are listed for comparison.

The results of this scan of various possible designs

```
# OF ELEMENT $ P-P RIPPLE IN PASSBAND (DB)? 4,.25
INPUT FILTER CENTER FREQ.(GHZ),BW(MHZ)&LOAD IMPEDENCE ZO? .440,3,50
INPUT GROUND PLANE SPACING , ROD DIAMETER
& DISTANCE TO CENTER OF FIRST AND LAST ROD? 1,.38..5
NO. OF FREQ. REJECTION PTS AND STEP SIZE (MHZ)? 38,.5
DESIGN DATA FOR 4 POLE INTERDIGITAL FILTER .BAND PASS RIPPLE .25 DB
CENTER FREQ.  .44 GHZ
CUTOFF FREQ.  .4385  (GHZ) AND  .4415   GHZ
RIPPLE BW.    3.000021E-03 GHZ
3 DB BW.      3.419328E-03 GHZ
FRACTIONAL BW. 6.81823E-03
FILTER Q    128.6803
EST QU      1459.315
LOSS BASED ON THIS QU   2.422713   DB
DELAY AT BAND CENTER  294.6652 NANOSECONDS
FREQUENCY REJECTION INFORMATION
                                                                * .4305  69
                                                                * .431   67
                                                                * .4315  65
                                                              *   .432   63
                                                            *     .4325  61
                                                         *        .433   58
                                                       *          .4335  56
                                                    *             .434   53
                                                 *                .4345  50
                                             *                    .435   46
                                         *                        .4355  42
                                     *                            .436   38
                                *                                 .4365  33
                          *                                       .437   27
                  *                                               .4375  19
        *                                                         .438   9
*                                                                 .4385  0
*                                                                 .439   0
*                                                                 .4395  0
*                                                                 .44    0
*                                                                 .4405  0
*                                                                 .441   0
*                                                                 .4415  0
        *                                                         .442   9
                  *                                               .4425  19
                          *                                       .443   27
                                *                                 .4435  33
                                     *                            .444   38
                                         *                        .4445  42
                                             *                    .445   46
                                                 *                .4455  50
                                                    *             .446   53
                                                       *          .4465  56
                                                         *        .447   58
                                                            *     .4475  61
                                                              *   .448   63
                                                                * .4485  65
                                                                * .449   67
QUARTER WAVELENGTH = 6.706136 INCHES
THE LENGTH OF INTERIOR ELEMENTS = 6.418164  INCHES
LENGTH OF END ELEMENTS = 6.438183  INCHES
GROUND-PLANE SPACE = 1       INCHES
ROD DIAMETER  = .38  INCHES
END PLATES .5  INCHES FROM C/L OF END ROD
TAP EXTERNAL LINES UP  .2294638  INCHES FROM SHORTED END
LINE IMPEDANCES: END ROD 67.31341  ,OTHER  72.49873  , EXT. LINES  50 OHM
DIMENSIONS
EL. NO.       END TO C       C TO C      G(K)         Q/COUP
0                                        1             1.570873
1               .5                       1.378239      .6633376
                            1.894495
 2              2.394495                 1.269327      .5431323
                            1.969942
 3              4.364437                 2.055808      .6633376
                            1.894495
 4              6.258931                 .8509719      1.570873
 5                                       1
                6.758931
```

fig. 4. Computer printout for the 440-MHz filter.

quickly reveals some expected trends. For instance, it is clear that filters that are too narrow (i.e., filters that have passband widths well under 1 percent of their center frequency) have increasingly higher passband losses. Also, filters with wide passbands have lower loss at the center frequency but decreased out-of-band rejection. These are fundamental tradeoffs. Next, it is also apparent that increasing the number of elements increases both the passband loss and the desired rejection of out-of-band signals, and that a filter with a relatively wide passband and many elements will have low in-band losses and good rejection. However, the number of elements cannot be increased arbitrarily because experience has shown that filters with many elements are less easily built and tuned, and that for Amateur construction and tuning techniques it is best to avoid the use of more than about five or six elements.

With all this in mind, we selected a four-section filter with passband ripple of 0.25 dB (VSWR = 1.62) and a bandwidth of 3 MHz. For our application, these choices resulted in approximately the desired loss and rejections, namely 2.4 dB loss at 440 MHz and a rejection of about 46.8 dB at 445 MHz. Note that this rejection is relative to the passband loss, so that the filter's total loss at 445 MHz is estimated to be approximately 49.2 dB.

computer output

The computer program's output for this filter design is shown in **fig. 4**. The printout contains information on both the electrical performance estimates as well as mechanical information in detail sufficient to fully describe the filter.

After the "RUN" command is entered, the program asks for some electrical design information such as the number of sections, or elements, in the filter and the passband ripple in decibels. Enter these two numbers, separated by a comma, followed by the "return" key, and the computer will respond with the next questions. These are the center frequency, expressed in gigahertz, the ripple bandwidth of the filter in megahertz, and the desired load impedance in ohms (usually 50). As before, these three entries should be separated by commas and followed by a return.

Next, the program requests some mechanical information. The spacing between the top and bottom ground planes, the diameter of the resonant rods and the desired spacing between the end of the filter and the first rod are entered in response to the questions. All of these dimensions should be entered in decimal inches because the design formulas within the program contain constants in inches.

The third and last section of input data concerns the plotter output. The operator must specify how many plot points and the spacing in megahertz between each point. The maximum number of plot points is 40. It is usually convenient to specify a step size of somewhat smaller than the passband width to give a good picture. In this example, the filter is 3 MHz wide and the plot with a one-half MHz resolution shows the rejection skirts clearly.

After these last bits of information are entered, the computer program proceeds without further intervention. It first prints out some of the calculated parameters of the filter. Then the computer graphically plots the pass and reject bands on the screen and on a printer if one is selected.

The last block of computer printout gives the mechanical details of the filter. The quarter wavelength listed is the inside dimension of the filter cavity. The length of the interior elements is listed, followed by the length of the two end elements. All of the interior elements have the same length, but the two end elements may have a different length from the end rods. The tap point is the point on the end element at which the external connection is made, and it is measured from the "cold" or grounded end of the rods. A tabular summary of the filter's dimensions appears at the end of the printout. It lists the end to center dimensions, the element center to center dimensions, and two coupling coefficients. The mechanical dimensions are easily translated into a sketch of the filter. Naturally, the designer must have certain dimensions and construction materals in mind before the computer program can be run. The ground plane spacing and rod diameter must be entered as constants to permit the completion of the design. These variables depend on the materials used to build the filter, and on the construction technique.

construction

Now that the computer program has been used to select an optimized filter and it has printed the mechanical dimensions of the filter's structure, it is time to consider how to translate the filter design into a form that can be realized. The mechanical structure must be sufficiently sound to produce the expected performance.

The table of numbers at the end of the computer printout, **fig. 4**, gives fairly complete information on the dimensions of the filter structure. The first column gives the element number, with zero indicating the first edge of the filter and a number one greater than the total number of rods denoting the other edge of the structure. The second column gives dimensions in inches from the end of the housing to the center of the rod indicated. For example, in our design, the second element is to be located a distance of 2.39 inches from the end of the housing. The total length of the filter will be the dimension listed opposite the final entry which is, in this case, 6.759 inches (17.168 cm).

Although the printout gives information on how to size the filter, it is probably a good idea to make a drawing to fix in your mind just how the box is to be assembled. **Figure 5** is a mechanical sketch of the 440 MHz filter. It is clear that the dimensions on this drawing have been taken directly from the computer printout, but the sketch also shows how the walls of the box are to be joined together. The dimensions on the printout are for the electrical housing, which consists of the copper conductor inside the box, and not the outside, mechanical dimensions. Keep in mind that the external dimensions of the box are not the critical factors.

One of the more popular amateur construction materials is copper clad printed circuit board. It is widely available at low cost, is strong and stable, and it can be easily joined together by regular soldering techniques to make boxes and circuit housings of any given size. It can be cut with a sheet metal shear or tin snips to fairly close tolerances, and so it is a good choice as the basic "building block" material for a custom bandpass filter. Furthermore, copper is one of the best conductors for use at high frequencies as its electrical resistivity is quite low. Of the more common materials, only silver has better high-frequency conductivity, and it is considerably more expensive.

To construct the filter housing of this example, we used 1/16 inch double-clad fiberglass epoxy board throughout. The top, bottom, sides and end pieces of the filter's housing were cut to shape with a sheet metal shear, drilled to accept the resonator rods, and soldered together with "tack" joints, that is with small flows of solder at intervals along the edges of the pieces to be joined. The pieces need not be soldered with a continuous seam, but a soldered tack should be used about every inch. This dimension corresponds to only about 5 percent of a wavelength, and so it ensures good electrical interconnection at all points along the copper.

The most critical dimensions of the housing are the width of the filter, which is usually a quarter wavelength, and the inside height of the cavity. A small piece of wood or metal with perpendicular faces is a good "jig" to help solder the housing walls accurately. If you work carefully, you should be able to build a housing that's both nearly square and accurate enough to take advantage of the custom design made possible with the computerized design aid described in this article.

Because we want to hold all of the parts of this filter together with regular tin-lead solder, all of the parts obviously should be solderable. A good material for the filter rods is common copper tubing which is available in a number of sizes in hardware and plumbing supply stores. For the design of this 440 MHz filter we chose 3/8 inch tubing. The "3/8 inch" refers to the

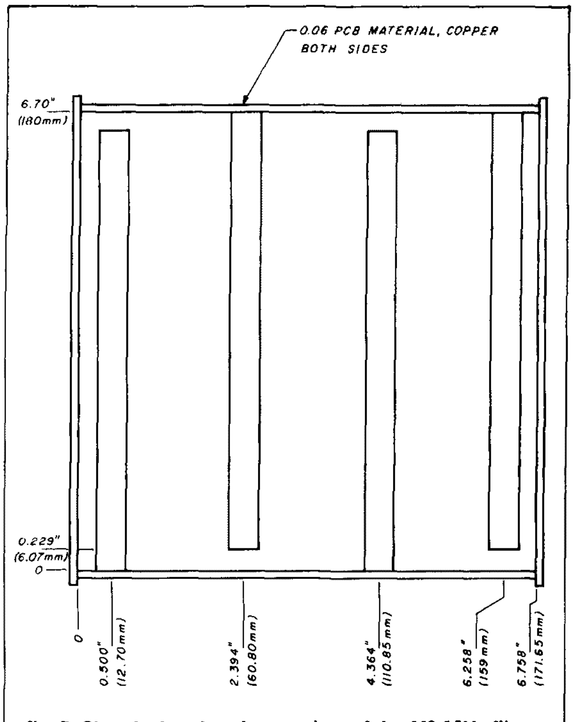

fig. 5. Sketch showing the top view of the 440-MHz filter. The end walls are deliberately cut longer so they can be soldered to both sides of the side walls.

outside dimension of the tubing, so the filter rod diameter measures approximately 0.375 inches.

The choice of rod diameter is not entirely arbitrary, although it is not very critical, either. As a rule of thumb, the rod diameter should be roughly 1/3 the housing height. In general, large diameter rods have lower losses than smaller rods. This is because skin effect losses predominate at radio frequencies. A larger diameter rod has greater surface area and, hence, less resistive loss than a smaller diameter rod. However, using a larger rod diameter can lead to mechanical difficulties. In this filter, for example, it was difficult to solder the large 3/8 inch diameter copper rods. These rods have fairly large masses and good thermal conductivity, so a regular soldering iron, intended for lighter duty circuit board work, just wouldn't provide enough heat. In the end, it took the greater heat of a propane torch to solder the rods to the copper walls.

The rods should be cut a bit longer than the correct length given by the printout. Then they are fit through holes drilled in the housing wall and soldered on both sides of the double clad circuit board, as shown in **fig. 6**. A good solder joint on each side increases the mechanical strength and improves grounding. The interior lengths of the rods and the center-to-center spacings between the rods are among the most critical dimensions affecting the frequency re-

sponse of the filter, so measure them as carefully and precisely as possible. In spite of all the care you use in construction, however, it will almost certainly be necessary to peak tune the filter response.

One simple way to tune the filter rod lengths is to load their open-circuited ends with variable capacitors. If the rods are just a bit shorter than the design calls for, then the small capacitance of a tuning screw at the open end will tune that element's resonant frequency. This tuning compensates for the minor inaccuracies which inevitably occur in construction.

These tuning screws need good grounds at the points where they penetrate the wall. It is important to realize that it is the inside grounded surface that matters most, because the inside copper cladding forms the conductive boundary that contains the filter's electric fields. For this reason the tuning screws are supported by brass nuts soldered to the inside surface of the box. These nuts are visible in the overall photograph of the filter and especially in the close-up view, **fig. 7**. The nut makes a simple threaded support for the screw and serves as the low impedance path from the screw body to the ground plane. On the outside of the filter housing a second nut holds the screw firmly in place. This nut is tightened after all of the filter tuning is completed, and it ensures that the screw is tightly bound to the soldered-down nut and prevents it from moving and thereby detuning the filter.

The computer program also lists the tap point distance, which is the position at which the external connectors pass through the walls and are coupled to the filter. The distance is given relative to the shorted end of the rods. This junction should be by a short length of wire or cable. The closeup view shown in **fig. 7** shows how the connector's center pin has been joined to the end rod with a short length of solid wire. The actual tap point distances may need to be adjusted for best performance, but if the design value is used it will serve for most cases without change.

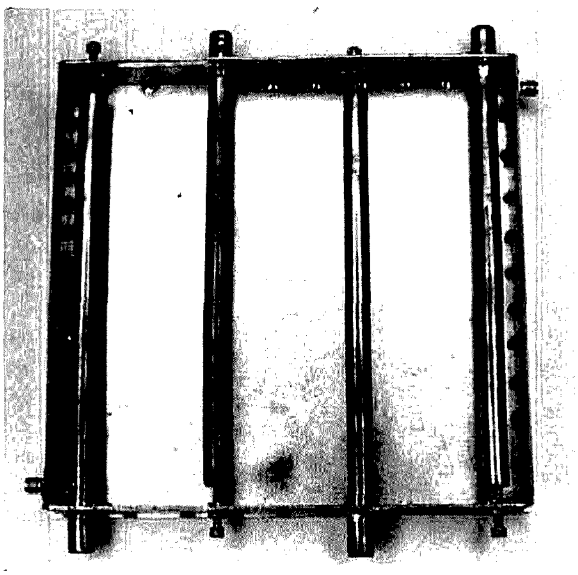

fig. 6. An overall view of the 4 section 440-MHz bandpass filter. The top cover has been removed to show the internal detail.

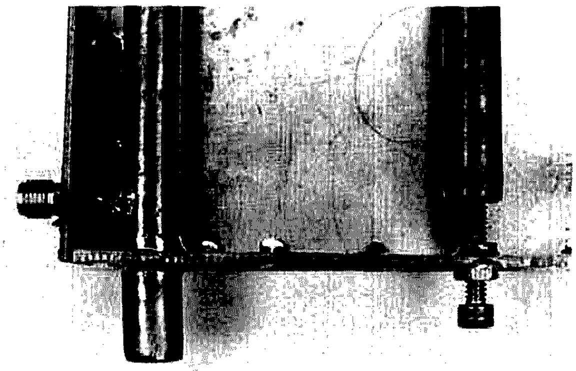

fig. 7. Close-up view of the 440-MHz filter showing details of the 50-ohm coaxial connector tap and of a tuning screw.

tuning

After the filter has been fully assembled, which means after all of the soldered joints are fully completed and the top is well secured electrically to the sides, it is time to test and tune.

Tuning microwave filters can be done in a number of ways. Several methods are described in reference 1, but the simple procedure of "sight tuning," or tuning by eye, is adequate for amateur filters, especially for filters with fewer than five or six sections. The basic principle involved is to tune for maximum signal at the center frequency, a process sometimes called synchronous tuning. The tuning is interactive, which means that one adjustment affects the tuning of adjacent elements, so it is necessary to return to each element once or twice to achieve peak performance.

A basic test set that can be used to peak the performance of any of the filters described here is shown in **fig. 8**. Inject a signal at the center frequency of the filter, detect or monitor the output power level, and tune for maximum. If the center frequency is within the ham bands, the input signal can be a transmitter or exciter. A signal generator can also be used. Harmonics of lower frequency sources or crystal oscillators can be useful as well. For example, the third harmonic of a 2-meter transmitter falls within the 420 to 450 MHz band and could be used to tune this filter. The signal detector can be a receiver with a signal level meter, a diode detector, a sensitive power meter or signal analyzer. Reference 6 describes many good low-cost UHF test methods in detail.

fig. 8. Simple test set block diagram. The signal source produces an output at the filter's center frequency. The filter is then tuned for maximum signal at the detector.

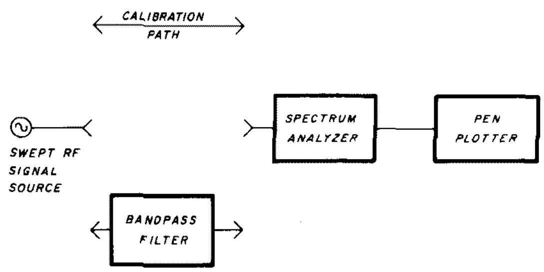

fig. 9. Test set used to produce the filter frequency response charts.

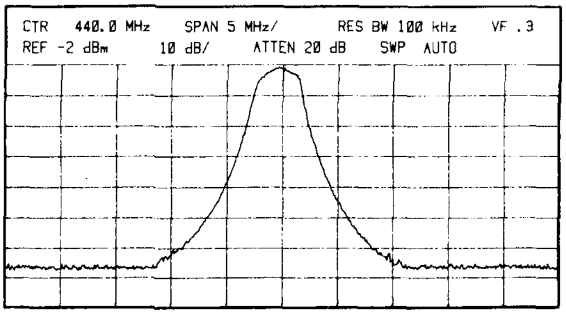

fig. 10. Swept frequency response of the four section 440-MHz filter. The horizontal scale is 5 MHz per division and the vertical scale is 10 dB per division. The spectrum analyzer settings were recorded automatically at the top of the plot.

Apply the test signal to one of the filter's connectors and connect the signal detector to the other and tune each of the screws to achieve the maximum output signal. The tuning range of this type of filter design is rather limited, which helps prevent tuning to the wrong harmonic, as can happen with broadly resonant circuits. This is a useful feature if the simple test set, with its potentially low spectral purity, is to be used.

If the filter is properly designed and carefully constructed, the slight tuning range afforded by the tuning screws should be sufficient. Each tuning screw should show a definite maximum. If it does not, this is an indication that the element which you are tuning is not properly resonant. In such a case, the rod length must be corrected before the filter will operate properly.

Once you've adjusted each tuning screw to yield the maximum output signal, go back and readjust each screw again slightly to peak the filter. If the filter response is not as calculated, and if the passband losses seem high even though each of the resonators gives a good peak tuning point, it may be necessary to adjust the tap points at the two end resonators. If no means of carefully measuring the losses is available, it is probably better to stay with the calculated tap dimensions. At this point the tuning is done.

A more sophisticated measuring system is useful for measuring the actual performance of the filter. **Figure 9** is a diagram of the test set which produced the swept frequency response of the filter examples. The input signal from a sweep generator scans across the frequency range in regular sweeps. A spectrum analyzer measures the output signal from the bandpass filter, and provides a graphic plot of the filter's output signal across the frequency range swept by the signal generator. Because the power from the generator to the filter input is constant, the spectrum analyzer's display is a direct representation of the filter's attenuation at various frequencies.

Figure 10 is the response of the 440-MHz filter. This graph is the plotter output from the test set described above. The filter was tuned using the simple single-frequency method of peaking all adjustments for maximum signal at 440 MHz. With the aid of such sophisticated test equipment it is possible to tune for improved passband flatness or for more precise centering if desired, but this clearly was not necessary in this case.

In summary, the 440 MHz bandpass filter example fully met its design goals of low cost, simple construction and easy adjustment, and it produced the desired electrical performance. The filter is physically small enough and sufficiently sturdy to be used in a fixed base system, although mechanical improvements would be needed for mobile service. This general construction technique using copper clad board and tubular copper or brass resonators has been applied to successfully build filters for the amateur bands from 440 to 2304 MHz.

1296 MHz filter

The second example of the use of this program is a bandpass filter centered at 1296 MHz with a desired ripple bandwidth of 20 MHz, a passband ripple of 0.25 dB and rejection points of 35 dB and 50 dB specified. Using these data, a few iterations with the program revealed that a four-section filter would again meet the requirements.

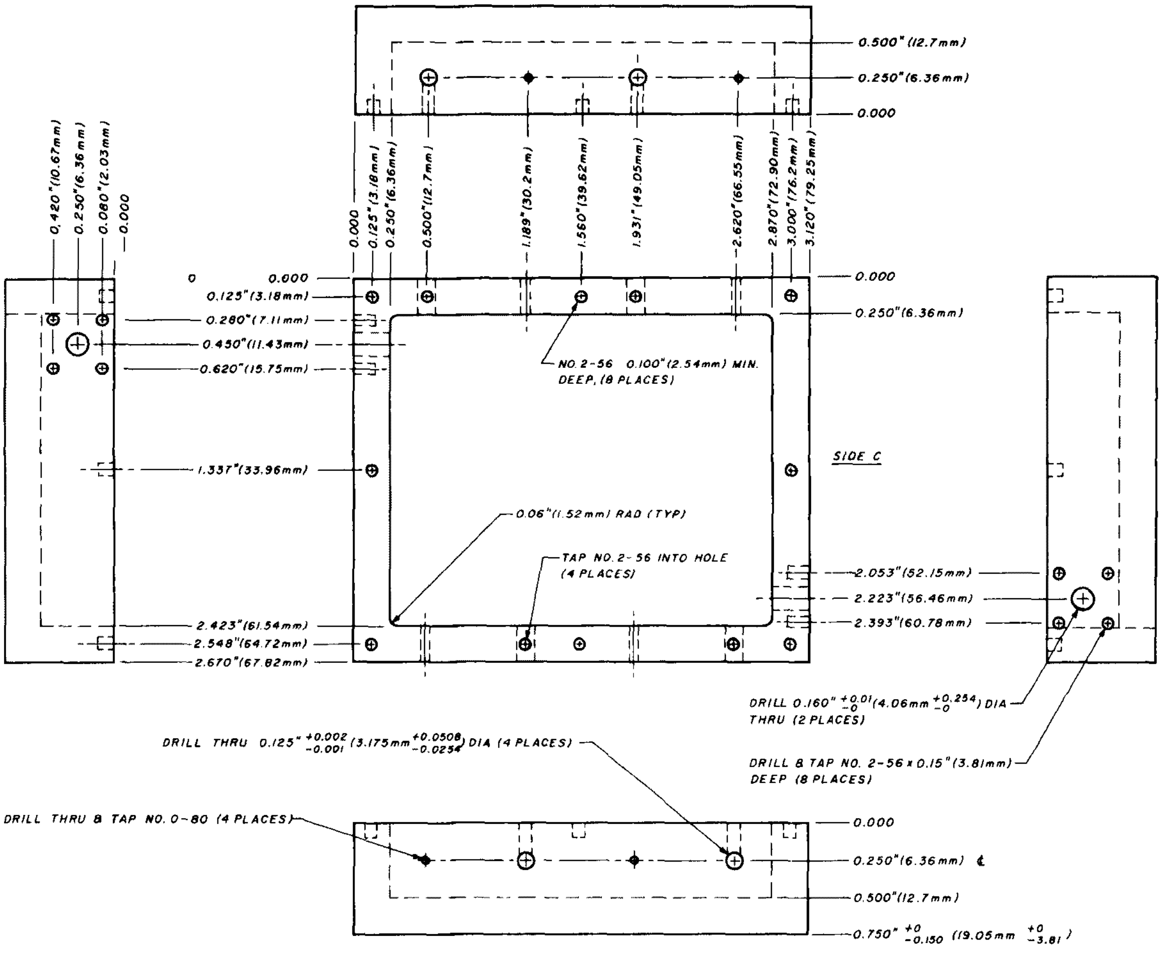

fig. 11. Mechanical details of the 1296-MHz filter housing. This housing is machined from aluminum stock. The top cover, not shown, is made from 0.06 sheet aluminum.

If the construction techniques used to build the 440-MHz filter seemed spartan, then this 1296 MHz filter is by contrast decidedly upscale. In order to prove that close tolerance construction could give good agreement with the computed data, a machined aluminum housing was used for this filter. Machining a housing using a metal milling machine gives precise control of the housing dimensions. This housing is considerably more accurate than the hand-made structure described in the first example.

The mechanical data supplied by the program were used to make a sketch, shown in **fig. 11**, of a housing that could be manufactured simply on a metal milling machine. The elements, machined from 0.125 inch brass rod stock (a standard size) were tightly "press fit" into holes in the housing walls. The rods are held tightly in position with a small setscrew once the exact interior lengths were determined. At the input and output ends of the filter SMA type connectors were installed so that their center pins contacted the rods at the tap point calculated by the computer program. Small diameter tuning screws were installed in the threaded holes in the walls opposite the open circuit end of each resonator rod. These screws, as in the 440-MHz filter, make it possible to fine-tune the filter passband. A top cover of 0.06-inch aluminum sheet was attached by eight screws that go into the threaded holes along the top edge of the housing. The photographs of this filter, **figs. 12** and **13**, show the construction details clearly. (Note: while 0-80 screws are specified in the construction drawings, 2-56 screws could be used as well.)

The housing was manufactured by machinists in a small shop, working from a simple sketch of the housing and cutting the aluminum stock with hand-operated machinery. In order to reduce costs, we

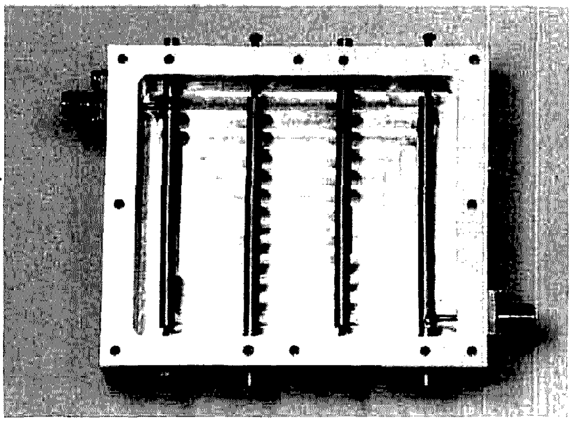

fig. 12. An overall view of the machined 1296-MHz filter. The sheet metal cover has been removed to show the internal cavity and the 4 resonator rods.

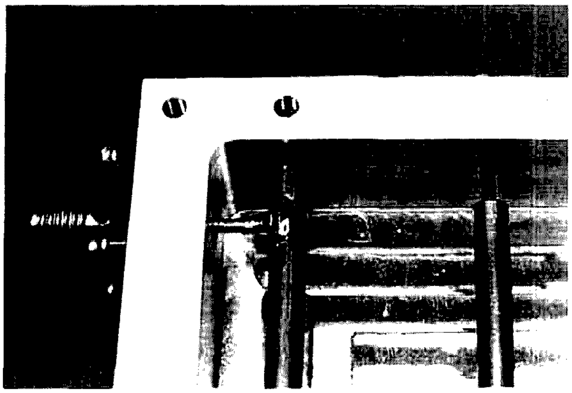

fig. 13. Close-up details of the 1296-MHz filter. The SMA connector center pin is clearly visible where it passes through the housing wall and contacts the end resonator rod. The small tuning screw at the end of the second resonator is also seen.

asked the shop to produce only a blank housing. They machined away the cavity to the precise 0.500 inch depth and drilled locating holes for the resonator rods, the tuning scews, the end connectors and the cover mounting screws, but they did not tap any of the holes. We were then able to finish the mechanical work by doing the time-consuming hand-tapping of all of the threaded holes. This reduced the cost of the housing by more than a third. Even so, the cost was in the $50.00 range, which may be justifiable only when precise results are essential.

However, the care and expense expended on the precise housing produced a filter that was nearly on frequency at first try, with precisely the initial resonator length settings that the computer predicted. Fine adjustments of the trimmer screws centered the passband precisely. **Figure 14** shows a plot of the swept RF response of the filter. Superimposed on the plot are circles that indicate the expected response calculated by the computer prediction. The calculated and actual values are in close agreement throughout the passband, and the rejection skirts are close to the computed values as well.

As before, the filter was peaked using the simple, single-frequency approach in order to illustrate the results obtainable with simple equipment. The test set used to make the swept frequency response plot was the same as that used in the 440 MHz filter tests.

This 1296 MHz filter, with its careful and precisely machined construction, shows the power and accuracy of the computer routine. The construction technique is a good one, and a simple metal milling machine, and perhaps even a drill press, can be used to make housings such as this one.

The use of 0.125 inch brass stock for the resonator rods was a bit of a compromise. It would have been better to use a somewhat larger diameter rod to reduce skin effect losses, but the 0.125-inch stock was on hand. Also, although brass rod is a good choice from a mechanical viewpoint, it is not a very good conductor of RF energy because its resistivity is about four times worse than copper's. Aluminum rods would be better than brass, because aluminum is both stronger and a better conductor, but with aluminum rods the tap point connections couldn't be easily soldered. Tradeoffs, as always, seemed to abound.

conclusion

This program is a powerful tool that greatly simplifies the selection and design of bandpass filters. The interdigital structure is useful from UHF to microwave frequencies, and provides good selectivity, low loss, small size, and an ease of construction that makes it suitable for many applications. The ease with which many different designs can be evaluated in software means that Amateurs can custom-design filters for specific applications and need not merely copy published designs that only approximate their re-

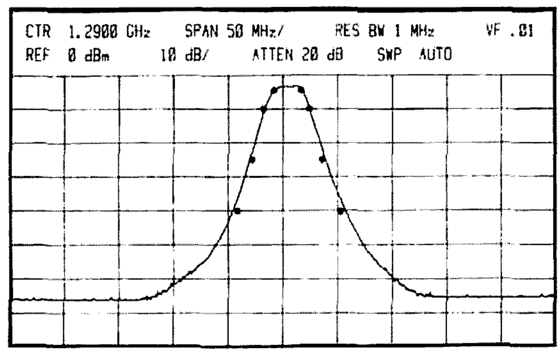

fig. 14. Frequency response curve of the 1296-MHz four section bandpass filter. The circles indicate the response computed by the program.

quirements. The widespread use of home computers together with software written specifically for Radio Amateurs should make possible a new generation of home-built equipment designs.

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ham radio

basic gamma matching

Simplify antenna matching with this BASIC program and your microcomputer

Antenna homebrewers have found the gamma match to be an ideal choice for matching a coaxial feedline to an all-metal radiator. It is simple to build, adds little weight and wind loading, is very strong mechanically, and allows you to match an unbalanced transmission line to either an unbalanced or balanced antenna.

Unfortunately, it isn't always easy to obtain a good match, and many have aborted their attempts in frustration. The problem is that initial gamma match dimensions are generally chosen arbitrarily. Sometimes you may be lucky and choose a reasonable starting point, but just as often your initial dimensions won't even be close. In this case you may spend hours on the tower going in circles looking for a match.

Formulas that allow you to generate gamma match designs are available.[1] However, the math involved is tedious, especially if several iterations must be performed. A home computer can simplify these calculations, allowing a variety of gamma matching networks to be examined in just a few minutes. A program that will design a gamma match for practically any Yagi or vertical antenna is presented here. While designed for the Apple II+, the program will work equally well with any microcomputer with only a few modifications.

background

The design of the gamma match is represented in **fig. 1**, and the schematic of the equivalent electrical circuit[2] in **fig. 2**. The circuit consists of a gamma rod and a resonating capacitor. The gamma rod of diameter d and length L runs parallel to the driven element of diameter D, separated by the center to center spacing S. It provides the desired resistance transformation, but at the same time introduces inductive reactance at the feed point. The gamma capacitor compensates for the inductive reactance, leaving only a resistive component.

Any of several gamma capacitors may be used. An air variable with adequate plate spacing for the anticipated power level is the usual choice. It may be mounted in a small weather resistant enclosure and connected to the gamma rod by means of a feedthrough insulator. Another method is to construct a coaxial capacitor within the gamma rod. This technique has been successfully applied by many commercial manufacturers and basement engineers.[3]

The shorting bar that determines length *L* generally takes the form of a strap, bent to conform to the driven element and gamma rod and secured with screws. When very long gamma rods are required, as when shunt feeding towers on the lower frequencies, a wire may be used for the gamma rod. If the diameter

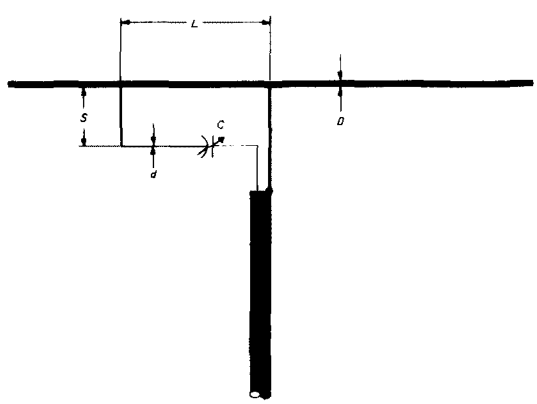

fig. 1. A representation of the mechanical design of the gamma match.

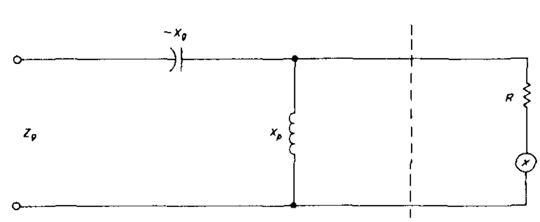

fig. 2. Schematic diagram of the gamma match. The antenna impedance R + jX is transformed to the feedline resistance Z_g by the gamma rod X_p and gamma capacitor $-X_g$.

By Richard A. Nelson, WB0IKN, Analog Technology, P.O. Box 8964, Fort Collins, Colorado 80525

```
10  HOME : CLEAR
20  PRINT "GAMMA MATCH DESIGN"
30  PRINT "BY RICHARD A. NELSON - WB0IKN"
40  PRINT
50  DEF  FN CSH(X) =  LOG (X +  SQR (X * X - 1))
60  PRINT "ENTER <M> FOR MONOPOLE"
70  INPUT "ENTER <D> FOR DIPOLE  > ";DM$
80  IF DM$ = "D" OR DM$ = "M" THEN  GOTO 100
90  GOTO 60
100  INPUT "ENTER FREQ IN MHZ > ";F
110  INPUT "ENTER FEEDPOINT RESISTANCE > ";RA
120  IF DM$ = "D" THEN RA = RA / 2
130  INPUT "ENTER FEEDPOINT REACTANCE > ";XA
140  IF DM$ = "D" THEN XA = XA / 2
150  INPUT "ENTER FEEDLINE RESISTANCE > ";RO
160  PRINT : PRINT
170  PRINT "(THE FOLLOWING ARE IN INCHES)"
180  PRINT : INPUT "ENTER DRIVEN ELEMENT DIAMETER > ";DE
190  INPUT "ENTER GAMMA ROD DIAMETER > ";DG
200  INPUT "ENTER GAMMA ROD SPACING > ";S
210 HZ = (1 + (( FN CSH((4 * S * S - DE * DE + DG * DG) / (4 * S * DG)))
     / ( FN CSH((4 * S * S + DE * DE - DG * DG) / (4 * S * DE))))) ^ 2
220 ZO = 60 *  FN CSH((4 * S * S - DE * DE - DG * DG) / (2 * DE * DG))
230 T = HZ / ZO
240 A = ((RO * XA) / (HZ * RA - RO))
250 B = (RO * ((RA) ^ 2 + (XA) ^ 2)) / (HZ * RA - RO)
260 Q = A +  SQR (A * A + B)
270 XS = HZ * ((RO * XA +  SQR ((RO * XA) ^ 2 + RO * (HZ * RA - RO) * ((
     RA) ^ 2 + (XA) ^ 2))) / (HZ * RA - RO))
280 LDGA =  ATN (Q * T)
290 LDG = (LDGA * 360) / (2 * 3.14159)
300 E = (RO / RA) * (((RA) ^ 2 + (XA) ^ 2) / Q)
310 G = (RO / RA) * XA
320 CR = 1000000 / (2 * 3.14159 * (E + G) * F)
330  HOME
340  IF DM$ = "D" THEN RA = RA * 2: IF DM$ = "D" THEN XA = XA * 2
350  PRINT
360  IF DM$ = "D" THEN  PRINT "DIPOLE ANTENNA"
370  IF DM$ = "M" THEN  PRINT "MONOPOLE ANTENNA"
380  PRINT
390  PRINT "FREQUENCY (MHZ) = ";F
400  PRINT "DRIVEN ELEMENT DIAM = ";DE
410  PRINT "GAMMA ROD DIAM = ";DG
420  PRINT "GAMMA ROD SPACING = ";S
430  PRINT "DRIVEN ELEMENT RESISTANCE = ";RA
440  PRINT "DRIVEN ELEMENT REACTANCE = ";XA
450  PRINT "FEEDLINE RESISTANCE = ";RO
460  PRINT
470  PRINT "GAMMA LENGTH (DEGREES) > ";LDG
480 FT = (948 / F) * (LDG / 360): PRINT "GAMMA LENGTH (FEET) > ";FT
490 IN = FT * 12: PRINT "GAMMA LENGTH (IN) > ";IN
500 CM = IN * 2.54: PRINT "GAMMA LENGTH (CM) > ";CM
510  PRINT "GAMMA CAP IN PF > ";CR
520  PRINT : PRINT "TYPE ANY KEY TO CONTINUE >": GET T$: GOTO 10
```

fig. 3. A listing of the program for an Apple II+ computer. Simple modifications will allow the program to be used with practically any home computer.

```
GAMMA MATCH DESIGN
BY RICHARD A. NELSON - WB0IKN

ENTER <M> FOR MONOPOLE
ENTER <D> FOR DIPOLE  > D
ENTER FREQ IN MHZ > 14.200
ENTER FEEDPOINT RESISTANCE > 20.0
ENTER FEEDPOINT REACTANCE > +7.5
ENTER FEEDLINE RESISTANCE > 50


(THE FOLLOWING ARE IN INCHES)

ENTER DRIVEN ELEMENT DIAMETER > 1.5
ENTER GAMMA ROD DIAMETER > .25
ENTER GAMMA ROD SPACING > 3.0

DIPOLE ANTENNA

FREQUENCY (MHZ) = 14.2
DRIVEN ELEMENT DIAM = 1.5
GAMMA ROD DIAM = .25
GAMMA ROD SPACING = 3
DRIVEN ELEMENT RESISTANCE = 20
DRIVEN ELEMENT REACTANCE = 7.5
FEEDLINE RESISTANCE = 50

GAMMA LENGTH (DEGREES) > 28.7984588
GAMMA LENGTH (FEET) > 5.34055927
GAMMA LENGTH (IN) > 64.0867112
GAMMA LENGTH (CM) > 162.780246
GAMMA CAP IN PF > 188.92213

TYPE ANY KEY TO CONTINUE >
```

fig. 4. The screen display for a sample run showing the sequence of data entry and output.

```
DIPOLE ANTENNA

FREQUENCY (MHZ) = 14.2
DRIVEN ELEMENT DIAM = 1.5
GAMMA ROD DIAM = .25
GAMMA ROD SPACING = 6
DRIVEN ELEMENT RESISTANCE = 20
DRIVEN ELEMENT REACTANCE = 7.5
FEEDLINE RESISTANCE = 50

GAMMA LENGTH (DEGREES) > 25.3194033
GAMMA LENGTH (FEET) > 4.69538231
GAMMA LENGTH (IN) > 56.3445877
GAMMA LENGTH (CM) > 143.115253
GAMMA CAP IN PF > 241.894184


DIPOLE ANTENNA

FREQUENCY (MHZ) = 14.2
DRIVEN ELEMENT DIAM = 1.5
GAMMA ROD DIAM = .5
GAMMA ROD SPACING = 3
DRIVEN ELEMENT RESISTANCE = 20
DRIVEN ELEMENT REACTANCE = 7.5
FEEDLINE RESISTANCE = 50

GAMMA LENGTH (DEGREES) > 38.3406199
GAMMA LENGTH (FEET) > 7.11011495
GAMMA LENGTH (IN) > 85.3213794
GAMMA LENGTH (CM) > 216.716304
GAMMA CAP IN PF > 262.100816
```

fig. 5. Calculated results for two additional runs on the antenna in *fig. 4*.

is too small to provide a match, a "cage" may be constructed, using several wires to effectively create a cylinder at the required radius. Variations in gamma match design are limited only by the mechanical and electrical integrity of the structure. This versatility greatly adds to the gamma's usefulness.

about the program

A listing of the program for the Apple II + computer is shown in **fig. 3**. Although written specifically for the Apple, I tried to keep the number of commands unique to Applesoft to a minimum. It should be easy for owners of other brands of microcomputers to translate the program for use on their machine.

Lines 10 through **40** clear the memory and the screen, and print the heading. **Line 50** defines the inverse hyperbolic cosine function needed in the calculations. **Lines 60** through **150** input data regarding the antenna and feedline, and **lines 180** through **200** input the driven element diameter and anticipated gamma match constants (gamma rod diameter and spacing). **Lines 210** through **320** perform the calculations of gamma rod length and capacitance. Note that since a gamma match essentially loads into only one side of a dipole, the feedpoint impedance of a dipole antenna must be divided in half before performing calculations. This is done by **lines 120** and **140. Line 340** restores the original impedance value before outputting data. **Line 330** clears the screen and **lines 360** through **510** output the specified and calculated data. **Line 520** returns the program to the beginning.

Once you have the program running, you may wish to add custom features. For example, you could change **line 520** to allow you to either return to the beginning, or to **line 180** if you want to evaluate a different gamma match design without changing the antenna parameters. If you'd like hard copy of the data, you could add the appropriate "printer on" and "printer off" commands between **lines 330** and **340**, and **510** and **520** respectively. For an Apple II the commands would look like this:

```
335 PR#1
515 PR#0
```

Another possibility is a "WILL NOT MATCH" message in response to a divide-by-zero error (for Apples, use ONERR GOTO), indicating that a different design should be tried. I included all of these in my

original program, but they were eliminated in the version presented here in order to simplify program entry and translation.

The hardest part of using this program will be determining the driven element impedance. Fortunately, the feedpoint characteristics of most common antennas are sufficiently documented to provide good data. The error introduced by "guesstimating" will in many cases be better than typical homebrew construction tolerances. If you have a noise bridge or impedance meter you should have no trouble determining the feedpoint impedance. Refer to any of the standard antenna texts if you are unsure about a particular antenna.

Once you have the impedance data — assuming that you know the diameter of your driven element and gamma rod — select an arbitrary gamma rod spacing. Start with an estimate based on mechanical and "eyeballed" electrical considerations. Three inches is a bit small for a shunt-fed tower, and a foot is obviously too big for a 2-meter Yagi.

Load and run the program. It will begin by asking whether you are evaluating a dipole or a monopole. Then it will request the values of a number of constants — feedpoint impedance, design frequency, etc. After all data has been entered, the computer will calculate and display the gamma rod length and the value of the gamma capacitor. If these values are not acceptable, run the program again, trying a different gamma rod spacing, and/or a different gamma rod diameter (if practical). You can vary the feedpoint impedance by slightly changing the length of the driven element. By examining a variety of alternative designs you can find the best combination, with reasonable mechanical and electrical parameters. If the antenna is not suitable for gamma matching, you can learn this quickly, without wasting hours in hands-on experimentation.

design examples

To demonstrate program operation, and to allow you to check operation once the program is keyed in, I will present several design examples.

Figure 4 shows the screen display on a run for a computer generated six-element, 20-meter Yagi design by Lawson.[4] The calculated feedpoint impedance of the antenna is 20 + j7.5 ohms. In this example, I've assumed 1.5-inch (38.1-mm) diameter element centers, a gamma rod diameter of 0.25 inch (6.35 mm), and a gamma rod spacing of 3 inches (76.2 mm). **Figure 5** shows the results of two additional gamma match designs for this antenna. In **fig. 5A** the gamma rod spacing has been increased to 6 inches (152.4 mm), and in **fig. 5B** the gamma rod diameter has been increased to 0.5 inch (12.7 mm).

Figure 6 shows the results for a monopole approximately 1/4-wavelength high. In this case the gamma rod is a length of No. 10 wire (approximate diameter = 0.1 inch or 2.54 mm). Run A is for a 60-foot (18.3-meter) tower used as vertical radiator on 3.8 MHz. Computer analysis shows its impedance to be approximately 33 + j1.3 ohms. Run B is for a 55-foot (16.8-meter) tower operated on the same frequency. The results show how a smaller gamma capacitor may be used if the radiator is made capacitively reactive by reducing its overall height.

```
MONOPOLE ANTENNA

FREQUENCY (MHZ) = 3.8
DRIVEN ELEMENT DIAM = 12
GAMMA ROD DIAM = .1
GAMMA ROD SPACING = 12
DRIVEN ELEMENT RESISTANCE = 33
DRIVEN ELEMENT REACTANCE = 1.3
FEEDLINE RESISTANCE = 50

GAMMA LENGTH (DEGREES) > 44.8174331
GAMMA LENGTH (FEET) > 31.0576949
GAMMA LENGTH (IN) > 372.692339
GAMMA LENGTH (CM) > 946.638541
GAMMA CAP IN PF > 122.616005


MONOPOLE ANTENNA

FREQUENCY (MHZ) = 3.8
DRIVEN ELEMENT DIAM = 12
GAMMA ROD DIAM = .1
GAMMA ROD SPACING = 12
DRIVEN ELEMENT RESISTANCE = 25
DRIVEN ELEMENT REACTANCE = -38
FEEDLINE RESISTANCE = 50

GAMMA LENGTH (DEGREES) > 53.7315208
GAMMA LENGTH (FEET) > 37.2350012
GAMMA LENGTH (IN) > 446.820015
GAMMA LENGTH (CM) > 1134.92284
GAMMA CAP IN PF > 76.9934529
```

fig. 6. Results of two runs on a 75-meter vertical monopole antenna.

I have used this program to design gamma matching networks for a 20-meter Yagi and a 2-meter cubical quad. In both cases the results were superb. No trimming of the gamma rod was necessary. Adjustment of the variable capacitors was all that was needed to achieve a 1:1 VSWR. The normal trial and error process was totally eliminated.

references

1. Harold F. Tolles, W7ITB, "How to Design Gamma Matching Networks," *ham radio*, May, 1973, page 46.
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3. Fred Reynolds, W2VS, "Simple Gamma Match Construction," *QST*, July, 1957, page 30.
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ham radio

determining basic moon coordinates

Track the moon with this computer program

Serious EME (Earth-Moon-Earth) communication requires a way of accurately determining the position of the moon. Of the several ways this can be done, moon-tracking by computer is one of the more convenient.

The program described here runs on TRS-80 models I/III with a minimum of 16K RAM and either level II/III or disk BASIC; it should run on other computers as well with only minimal changes. With your QTH as the point of reference, it provides readings of azimuth, elevation, GHA, declination, and right-ascension of the moon as well as relative path loss, local time of day and local sidereal time (LST). The positional data are expressed in *topocentric* values — i.e., they've been corrected for your location on the surface of the earth. (Many other moon tracking programs provide *geocentric* data referenced to the center of the earth.)

To output the above information, you simply input your latitude and longitude and the day(s) and time(s) in GMT (Greenwich Mean Time). The computer displays the information on the CRT monitor, and/or sends it to the printer. **Figure 1** shows a sample output for W2WD (40 degrees 39 minutes North, 74 degrees 22.5 minutes West) on June 8, 1983. Although this output came from the printer, the CRT display was similar; the data appears on the printer at the same time it's displayed on the CRT.

keyboard input data

Figure 2 shows a screen printout of the inputs required to produce the output data discussed above. After loading, RUN "MOON" <ENTER> starts the program from either level II or disk BASIC. Once initiated, the logo appears and suggests a listing of lines 1000-1360 if you need a short description of the program. If you hit <ENTER> to go on, the computer asks, WHAT ARE THE STATION CALL LETTERS? (As shown, I entered W2WD.) Next, the computer asks WHAT IS YOUR LOCAL TIME (EST, EDT, PST, etc.)? (The reply was EST.) and then, HOW MANY HOURS/MINUTES DIFFERENCE FROM GMT? USE + IF EARLIER, − IF LATER (e.g., EST would be −0500). (The answer was −0500.) Next it asks, YOUR LATITUDE (DEGREES, MINUTES) +NORTH/−SOUTH? (In my case this was 40,39 for 40 degrees and 39 minutes north latitude. If you are in the southern hemisphere, you must use a negative sign, e.g., −40,39.) Longitude is entered in a similar fashion. (My input was 74,22.5 for 74 degrees and 22.5 minutes west longitude.) Use a negative sign (−) for east longitude.

The accuracy of the final calculations depends on the precisely accurate determination of your geodetic location. Your town manager or city engineer may be able to help. If not, you can use maps sold by the U.S. Geological Survey, especially those drawn to 1:24,000 scale. Some automotible road maps may show sufficient detail to allow fairly close determination of your location.

The next information to be inputted is the DESIRED PRINTING INCREMENT IN MINUTES (1-60). I used 20 minutes between each line of output data, but this can be anything between 1 and 60 minutes. Positions can be determined for time intervals of less than 1 minute apart, but the displays and printouts, as the program is now written, will round the results to increments of full minutes — i.e., if you need data every 6 seconds (0.1 minutes), you can input 0.1 as the print-

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POSITION OF THE MOON ON 6 / 8 / 1983 GMT FROM W2WD

GMT	GHA	DEC	LST	EST	AZ	EL
0800	337.2	+10.5	2006	0300	77.4	1.4
0820	342.0	+10.5	2026	0320	80.5	5.1
0840	346.8	+10.6	2046	0340	83.5	8.7
0900	351.7	+10.7	2106	0400	86.5	12.5
0920	356.5	+10.8	2126	0420	89.6	16.2
0940	1.4	+10.9	2147	0440	92.7	19.9
1000	6.2	+10.9	2207	0500	95.9	23.6
1020	11.1	+11.0	2227	0520	99.2	27.4
1040	15.9	+11.1	2247	0540	102.7	31.0
1100	20.8	+11.2	2307	0600	106.4	34.6
1120	25.7	+11.3	2327	0620	110.4	38.2
1140	30.6	+11.3	2347	0640	114.7	41.7
1200	35.5	+11.4	0007	0700	119.4	45.0
1220	40.3	+11.5	0027	0720	124.6	48.2
1240	45.2	+11.6	0047	0740	130.3	51.2
1300	50.1	+11.6	0107	0800	136.8	54.0
1320	55.0	+11.7	0127	0820	144.1	56.4
1340	59.9	+11.8	0147	0840	152.1	58.5
1400	64.8	+11.9	0207	0900	161.0	60.0
1420	69.7	+11:9U	0227	0920	170.5	61.0
1440	74.6	+12.0U	0247	0940	180.4	61.4
1500	79.5	+12.1U	0307	1000	190.4	61.1
1520	84.4	+12.2U	0327	1020	200.0	60.2
1540	89.3	+12.2U	0348	1040	208.9	58.7
1600	94.2	+12.3U	0408	1100	217.1	56.7
1620	99.1	+12.3W	0428	1120	224.4	54.4
1640	103.9	+12.4W	0448	1140	231.0	51.7
1700	108.8	+12.5W	0508	1200	236.8	48.7
1720	113.7	+12.5W	0528	1220	242.1	45.6
1740	118.6	+12.6W	0548	1240	246.9	42.3
1800	123.5	+12.7W	0608	1300	251.3	38.9
1820	128.3	+12.7W	0628	1320	255.3	35.4
1840	133.2	+12.8J	0648	1340	259.1	31.8
1900	138.1	+12.8J	0708	1400	262.6	28.2
1920	142.9	+12.9J	0728	1420	266.0	24.6
1940	147.7	+12.9J	0748	1440	269.3	21.0
2000	152.6	+13.0J	0808	1500	272.5	17.3
2020	157.4	+13.1J	0828	1520	275.6	13.7
2040	162.3	+13.1	0848	1540	278.7	10.1
2100	167.1	+13.2	0908	1600	281.8	6.5
2120	171.9	+13.2	0928	1620	284.9	3.0

R.A. OF MOON = 0308 PATH-LOSS INCREASE + 1.3 DB

fig. 1. Moon coordinate printout for W2WD on 8 June 1983.

ing increment. You'll get ten different lines of data for every minute of readout, but the GMT, LST, and your local time will not change until the data for the following minute appears, etc. If this is an important use, you can modify the program to print out the time(s) to the resolution needed.

For EME operators who use ground reflection for added gain or cannot elevate their antennas, the pro-

```
WHAT ARE THE STATION CALL LETTERS ? W2WD
WHAT IS YOUR LOCAL TIME (EST, EDT, PST, ETC.) ? EST
HOW MANY HOURS/MINUTES DIFFERENCE FROM GMT?
    USE + IF EARLIER, - IF LATER
    (EG., EST WOULD BE -0500)? -0500
WHAT IS YOUR LATITUDE (DEGREES,MINUTES)
+ NORTH  /  - SOUTH  ? 40,39.0
WHAT IS YOUR LONGITUDE (DEGREES,MINUTES)
+ WEST  /  - EAST  ? 74,22.5
WHAT IS THE DESIRED PRINTING INCREMENT IN MINUTES (1-60)? 20
DO YOU ONLY WANT PRINTOUT WHEN THE MOON IS NEAR THE HORIZON
(YES/NO)? NO
DO YOU WANT HARDCOPY (YES/NO)? YES


INPUT - GMT MONTH, DAY, YEAR, TIME BEGINNING, TIME ENDING
USE 4-DIGITS FOR YEAR AND 24-HOUR CLOCK
ENTER DATA FOR UP TO 31 DAYS.
HIT <ENTER> AFTER LAST ENTRY

DATE 1  (MM,DD,YYYY,TTTT,TTTT) ? 6,8,1983,0,2400
DATE 2  (MM,DD,YYYY,TTTT,TTTT) ? 0_
```

fig. 2. Screen printout of input commands.

gram has an option to print or display data only when the moon is near the horizon. In this case you answer YES to DO YOU ONLY WANT PRINTOUT WHEN THE MOON IS NEAR THE HORIZON (YES/NO)? (I operate with a polar-mounted antenna so I answer NO to this question.) If you answer YES, you will be asked to reply to BELOW WHAT ELEVATION IN DEGREES DO YOU WANT PRINTOUT? You then enter the maximum elevation angle you're interested in.

If you don't have a printer or don't want hard copy, answer NO to DO YOU WANT HARDCOPY (YES/NO)? If you have a printer and answer YES but the printer is not turned on, the computer will reply PRINTER NOT READY. The program won't hang up, but will repeat the question until you either answer NO or turn on the printer.

At this point, you specify the date-time periods for the output data. These must be entered in GMT format. It is necessary to use *four* digits for the year; don't use 85, if you mean 1985! Similarly, time is inputted with four digits using a 24-hour clock for both the starting and ending points. You don't have to start the data at 0000 and end at 2400; they can be set according to your operating requirements. For the low-elevation-only option, you don't need to enter time spans because the computer will print only data meeting your criteria of when the elevation of the moon is below the angle you have specified earlier.

You can enter up to 31 dates for any given run. The dates do not have to be consecutive or in any given order, nor does the data requested have to be uniform from day to day. In fact, each entered date is a separate request. You may repeat the same dates with different start/end times if you wish.

program operation

Keep in mind that the calculations are relatively slow. Compiling data for 31 days could require hours of computer time. The actual time will vary greatly according to the time increments you specify, the printer speed, and other factors.

When you've completed the entry of all start/end times, hit <ENTER> to terminate the data input phase. The program prints the heading information and begins calculations. Each line of data requires roughly 15 seconds, even though the result of that calculation may be below the horizon and therefore not printed out. If you've asked for data every 20 minutes following midnight on a day when the moon does not rise until 0800 GMT, the computer will calculate the position of the moon 24 times before any results appear on the CRT or printer. Therefore, the program may seem to be hung up for several minutes even though it's actually hard at work calculating data that ends up below the horizon and is consequently not displayed.

fig. 3. Enhanced version moon coordinate program listing for TRS-80.

```
1000 '************* MOON COORDINATES ************
1010 '
1020 'PRIMARILY FOR USE IN EARTH-MOON-EARTH
1030 '(EME) COMMUNICATIONS BY RADIO AMATEURS.
1040 '
1050 'BASED ON PROGRAMS BY LANCE COLLISTER (WA1JXN/WA3GPL)
1060 'AND JAY LIEBMANN (K5JL).
1070 'SEE EIMAC PUBLICATIONS AS-49-6, AS-49-17 AND AS-49-24.
1080 'VARIAN, EIMAC DIVISION
1090 '301 INDUSTRIAL WAY
1100 'SAN CARLOS, CA 94070.
1110 '
1120 'MODIFIED FOR MODEL I TRS-80 LEVEL II AND DISK BASIC
1130 'WITH ENHANCED DISPLAYS AND ADDITIONS OF SIDEREAL TIME,
1140 'RIGHT ASCENSION AND DISTANCES TO THE MOON (CONVERTED
1150 'TO PATH-LOSS VARIATIONS IN DECIBELS)
1160 'BY WARREN BUTLER (W2WD).
1170 '
1180 'INPUT DATA: LATITUDE, LONGITUDE, GMT DATE/TIME AT SITE.
1190 'DATA FOR UP TO 31 DIFFERENT DAYS CAN BE INPUTTED AT
1200 'ONE TIME.  ENTER DATA IN THE FORMAT REQUESTED.  AFTER
1210 'THE LAST INPUT, INSERT ZEROS OR HIT <ENTER>.
1220 '
1230 'OUTPUT DATA: GHA, DECLINATION, AZIMUTH AND ELEVATION OF
1240 'MOON, SIDEREAL TIME (ST) AND LOCAL TIME, UNIVERSAL
1250 'WINDOWS FOR EME COMMUNICATION, RIGHT ASCENSION OF MOON,
1260 'PATH-LOSS VARIATIONS (DB).
1270 '
1280 'HARDCOPY OUTPUT CAN BE SELECTED IF PRINTER IS AVAILABLE.
1290 '
1300 'UNIVERSAL EME WINDOWS ARE SHOWN BY LETTERS FOLLOWING DEC.
1310 'U = EUROPEAN UNIVERSAL WINDOW
1320 'W = W/VE UNIVERSAL WINDOW
1330 'J = J/VK/ZL UNIVERSAL WINDOW
1340 '
1350 'BE PATIENT, THE CALCULATIONS CAN TAKE SEVERAL MINUTES.
1360 '
1370 PRINT
1380 PRINT
1390 CLEAR 500
1400 DIM F(31), V(31), Y(31), Q(31), S(31)
1410 P5=2.0000000000*3.1415926535
1420 D5=360.0000000000/P5
1430 R5=P5/360.0000000000
1440 CLS
1450 'GOTO 1730
1460 PRINT @ 0, STRING$(64, 170)
1470 PRINT @ 148, "MOON COORDINATE PROGRAM"
1480 PRINT @ 278, "WARREN BUTLER (W2WD)"
1490 PRINT @ 343,"2305 MORSE AVENUE"
1500 PRINT @ 404,"SCOTCH PLAINS, NJ 07076"
1510 PRINT @ 473,"(201) 233-4460"
1520 PRINT @ 576, STRING$(64, 170)
1530 PRINT @ 833,"IF YOU NEED A DESCRIPTION OF PROGRAM, LIST LINES
     1000-1360"
1540 PRINT @ 978, "ELSE HIT <ENTER> TO CONTINUE";
1550 INPUT I%
1560 CLS
1570 REM: BEGIN INPUT DATA SEQUENCE
1580 PRINT"WHAT ARE THE STATION CALL LETTERS ";
1590 INPUT W$
1600 INPUT"WHAT IS YOUR LOCAL TIME (EST, EDT, PST, ETC.) ";TL$
1610 PRINT"HOW MANY HOURS/MINUTES DIFFERENCE FROM GMT?"
1620 PRINT"    USE + IF EARLIER, - IF LATER"
1630 INPUT "    (EG., EST WOULD BE -0500)";TD
1640 IF TD<-1200 OR TD>1200 THEN 1610
1650 PRINT"WHAT IS YOUR LATITUDE (DEGREES,MINUTES) "
1660 PRINT"+ NORTH  /  - SOUTH  ";
1670 INPUT L5, U5
1680 IF L5>90 OR L5<-90 OR U5>60 THEN 1650
1690 PRINT"WHAT IS YOUR LONGITUDE (DEGREES,MINUTES) "
1700 PRINT"+ WEST   /  - EAST   ";
1710 INPUT L6, U6
1720 IF L6>180 OR L6<-180 OR U6>60 THEN 1690
1730 'W$="W2WD": L5=40: U5=39.0: L6=74: U6=22.5: TL$="EST": TD=-0500
1740 L5=(L5+U5/60)*R5
1750 L6=(L6+U6/60)*R5
1760 INPUT"WHAT IS THE DESIRED PRINTING INCREMENT IN MINUTES (1-60)";I
1770 IF I<=0 OR I>60 THEN 1760
1780 B$=""
1790 INPUT"DO YOU ONLY WANT PRINTOUT WHEN THE MOON IS NEAR THE HORIZON
      (YES/NO)";B$
```

(continued on page 43)

This happens in the example shown in **fig. 1**. On June 8, 1983, the moon rose at about 0800 GMT. The computer did not start printing out data until *six minutes* after the input phase had been completed and the <ENTER> key had been pressed. From that point the printing continued until moonset had been reached at about 2120 GMT. However, the program did not stop calculations until the specified ending time of 2400 GMT. The program required about 15 minutes to generate the data shown in **fig. 1**.

Similarly, if data ends above the maximum angle selected in the low-elevation mode, the program will also appear to be "hung-up." If you already know the times of moonrise and moonset, the start and end times can be set accordingly and much time can be saved in the printout process.

printouts and displays

Now let's look at the output data (**fig. 1**) in more detail. After the GMT time column, the GHA, or Greenwich Hour Angle, is shown. It is the angle subtended by the moon and the Greenwich meridian. This angle can be translated to your geographical meridian (Local Hour Angle) by adding your longitude if east, or subtracting if west. The third column provides the declination of the moon or its position north (+) or south (−) of the celestial equator. Using the GHA and declination information, the polar-mounted EME antenna can be kept trained on the moon.

Note that some of the values for declination are followed by the letter U, W, or J. These are indicators that the moon is in one of the universal EME windows established to allow use of fixed antenna arrays. This was intended to permit large antennas to be built at lower cost by eliminating the need to position the array in azimuth or elevation. The U, or universal window, was set up for contacts between European and USA stations. The W window is better situated for USA/USA and the J window is for JA/VK/W contacts.[4]

The next two columns show local time readouts. Standard or daylight-savings time needs no further discussion. LST (Local Sidereal Time) is tied to the stars and used primarily by astronomers. It is included here to help readers locate some of the stellar noise sources for checking EME systems performance. Or, for that matter, to locate a "cold" area of the sky for the same purpose. Many EME operators use the sun as a noise source to check antenna and system performance. When system performance is improved beyond the norm, smaller noise sources in the heavens such as Cassiopeia A, Cygnus A, and others can prove useful. Knowing the R.A. (right ascension of the moon) and declination of these sources and your LST can help in training your antenna to the proper direction in the sky.

For EME operators using AZ-EL antenna mounts, the computer calculates the azimuth and elevation angles to the moon from their QTH. Azimuth is the angle with respect to true north, (not magnetic north as would be indicated with a compass); elevation is the angle above the local horizon.

Below the positional data in **fig. 1** are two other useful parameters for EME operation. The first is the R.A., which indicates where the moon is in relation to other objects in the sky. Some of the stars are noise sources that can reduce the signal-to-noise ratio of the communication path when they are in line with the moon at the time of a QSO. Earlier, I suggested that these same kinds of noise sources could be useful for system calibration. For better EME contacts, however, they should be avoided. Reference 1 contains radio sky maps over the frequency range of 64 to 910 MHz. These show the noise temperature of the sky as a function of right ascension and declination. It should be mentioned that the right ascension of the moon given at the bottom of the printout listings is the first calculated value. The R.A. varies about one hour over a 24-hour day, but this usually provides sufficient accuracy to judge whether noise sources from behind the moon are going to be a problem.[2]

Because the moon is traveling in an elliptical path around the earth, its distance from the earth varies. At apogee it is roughly 407,000 kilometers from the earth; at perigee, 356,000 km. The variation in range is sufficient to add about 2 dB of additional path loss at apogee as compared to when the moon is at the perigee position. Apogee to perigee time spans are roughly 13 days. For each day of a printout, the path-loss difference in dB, as compared to the perigee position, is printed at the bottom of the listing. Equations used for calculating the earth-to-moon distances were taken from reference 3.

accuracies

Considering the number and types of calculations this program processes, it is reasonable to ask whether the results obtained on a personal computer are sufficiently accurate for use by the typical Amateur EME station. To answer this question, I compared the output data to the most accurate, reliable data available to me.

For GHA and declination I was unable to locate a suitable source of topocentric values to confirm the output of this program. However, the *Nautical Almanac* provides geocentric data with errors less than 0.0005 degrees and thus served as a satisfactory standard for data referenced to the center of the earth.[5,6] In order to compare "oranges to oranges," I had to bypass the translational calculations I had inserted to convert from geocentric to topocentric coordinates. I then used the random number generator of the

fig. 3, continued

```
1800 IF LEFT$(B$, 1)="Y" THEN 1830
1810 I6=100
1820 GOTO 2030
1830 INPUT"BELOW WHAT ELEVATION IN DEGREES DO YOU WANT PRINTOUT";I6
1840 WW$=" "
1850 INPUT"DO YOU WANT HARDCOPY PRINTOUT (YES/NO)";WW$
1860 IF LEFT$(WW$, 1)="Y" WW$="YES"
1870 IF WW$="YES" AND PEEK(14312) > 127 THEN PRINT "PRINTER NOT READY":
     WW$="": GOTO 1850
1880 PRINT
1890 PRINT"WHAT ARE THE GMT MONTH, DAY, YEAR DESIRED?"
1900 PRINT"*** NOTE - USE 4-DIGITS FOR YEAR (EG., 1983) ***"
1910 PRINT"ENTER DATA FOR UP TO 31 DAYS"
1920 PRINT"HIT <ENTER> AFTER LAST ENTRY"
1930 PRINT
1940 N=0
1950 FOR N=1 TO 31
1960 PRINT"DAY ";N;"  (MM,DD,YYYY)";
1970 INPUT F(N), V(N), Y(N)
1980 IF F(N)=0 THEN 2220
1990 IF F(N)<1 OR F(N)>12 OR V(N)<1 OR V(N)>31 OR Y(N)<1900 OR Y(N)>2000
     THEN 1960
2000 IF N=31 THEN 2220
2010 NEXT N
2020 GOTO 1950
2030 WW$=""
2040 INPUT"DO YOU WANT HARDCOPY (YES/NO)";WW$
2050 IF LEFT$(WW$, 1)="Y" WW$="YES"
2060 IF WW$="YES" AND PEEK(14312) > 127 THEN PRINT "PRINTER NOT READY":
     WW$="": GOTO 2040
2070 PRINT
2080 PRINT"INPUT - GMT MONTH, DAY, YEAR, TIME BEGINNING, TIME ENDING"
2090 PRINT"USE 4-DIGITS FOR YEAR AND 24-HOUR CLOCK"
2100 PRINT"ENTER DATA FOR UP TO 31 DAYS."
2110 PRINT"HIT <ENTER> AFTER LAST ENTRY
2120 PRINT
2130 N=0
2140 FOR N=1 TO 31
2150 PRINT"DATE";N;" (MM,DD,YYYY,TTTT,TTTT) ";
2160 INPUT F(N), V(N), Y(N), Q(N), S(N)
2170 IF F(N)=0 THEN 2220
2180 IF F(N)<1 OR F(N)>12 OR V(N)<1 OR V(N)>31 OR Y(N)<1900 OR Y(N)>2000
     OR  Q(N)<0 OR Q(N)>2359 OR S(N)<0001 OR S(N)>2400 THEN 2150
2190 IF N=31 THEN 2220
2200 NEXT N
2210 GOTO 2140
2220 IF N=31 THEN N5=N ELSE N5=N-1
2230 FOR N=1 TO N5
2240 IF LEFT$(B$, 1)="Y" THEN 2260
2250 GOTO 2290
2260 E1=2400
2270 B=0
2280 GOTO 2310
2290 E1=S(N)
2300 B=Q(N)
2310 M=F(N)
2320 D=V(N)
2330 Y=Y(N)
2340 Y1=Y-( INT(Y/100)*100)
2350 REM: SETUP OUTPUT FORMAT
2360 PRINT
2370 IF WW$="YES" THEN LPRINT""
2380 PRINT
2390 IF N=1 THEN CLS
2400 IF WW$="YES" THEN LPRINT""
2410 PRINT"POSITION OF THE MOON ON ";M;"/";D;"/";Y;" GMT FROM"" "
     W$
2420 IF WW$="YES" THEN LPRINT"POSITION OF THE MOON ON ";M;"/";D;"/";Y;"
     GMT FROM"" " W$
2430 PRINT
2440 IF WW$="YES" THEN LPRINT""
2450 PRINT"GMT" TAB(8)"GHA" TAB(17)"DEC" TAB(27)"LST" TAB(37)TL$
     TAB(47)"AZ" TAB(56)"EL"
2460 IF WW$="YES" THEN LPRINT"GMT" TAB(8)"GHA" TAB(17)"DEC" TAB(27)"LST"
     TAB(37)TL$ TAB(47)"AZ" TAB(56)"EL"
2470 I1=2
2480 REM: CALCULATION OF JULIAN DATE
2490 IF M>=3 THEN 2570
2500 IF INT((Y-1853)/4)<11 THEN 2530
2510 C1=-1
2520 GOTO 2540
2530 C1=0
2540 J1=365*(Y-1853)+D+30*(M+9)+INT((M+10)/2)
2550 J2=INT((Y-1853)/4)+1+C1
2560 GOTO 2680
2570 IF INT((Y-1852)/4)<11 THEN 2600
2580 C1=-1
2590 GOTO 2610
```

(continued on page 44)

fig. 3, continued

```
2600 C1=0
2610 IF M=9 THEN 2650
2620 IF M=11 THEN 2650
2630 C2=0
2640 GOTO 2660
2650 C2=1
2660 J1=365*(Y-1852)+D+30*(M-3)+INT((M-2)/2)
2670 J2=INT((Y-1852)/4)+C1+C2
2680 J=J1+J2
2690 JD#=J+2397547.5
2700 GOSUB 4450       'FOR MOON DISTANCE CALCULATIONS
2710 T1=J-17472.5
2720 REM: MAIN CALCULATIONS BEGIN
2730 D9=(B-INT(B/100)*100)+INT(B/100)*60
2740 D6=(E1-INT(E1/100)*100)+INT(E1/100)*60
2750 D7=D9-D6
2760 D8=D7-1
2770 IF D7>0 THEN 2790
2780 GOTO 2820
2790 IF D8>=0 THEN 4250
2800 B=E1
2810 REM: CALCULATION OF LATITUDE AND LONGITUDE OF MOON
2820 T=(B-INT(B/100)*100)/1440+INT(B/100)/24
2830 T5=T1+T
2840 K1=((.751213+.036601102*T5)-INT(.751213+.036601102*T5))* P5
2850 K2=((.822513+.0362916457*T5)-INT(.822513+.0362916457*T5))* P5
2860 K3=((.995766+.00273777852*T5)-INT(.995766+.00273777852*T5))
     *P5
2870 K4=((.974271+.0338631922*T5)-INT(.974271+.0338631922*T5))* P5
2880 K5=((.0312525+.0367481957*T5)-INT(.0312525+.0367481957*T5))
     *P5
2890 L8=K1+.658*R5*SIN(2*K4)+6.289*R5*SIN(K2)
2900 L8=L8-1.274*R5*SIN(K2-2*K4)-.186*R5*SIN(K3)
2910 L8=L8+.214*R5*SIN(2*K2)-.114*R5*SIN(2*K5)
2920 L8=L8-.059*R5*SIN(2*K2-2*K4)-.057*R5*SIN(K2+K3-2*K4)
2930 K6=K5+.6593*R5*SIN(2*K4)+6.2303*R5*SIN(K2)-1.272*R5* SIN(K2-2*K4)
2940 L7=5.144*R5*SIN(K6)-.146*R5*SIN(K5-2*K4)
2950 REM: CALCULATION OF RIGHT ASCENSION (R1) AND DECLINATION (D1)
2960 D1=COS(L7)*SIN(L8)*.397821+SIN(L7)*.917463
2970 D1=ATN(D1/( SQR(1-D1[2)))
2980 G1=50+.5+((D1)/(.792))*D5
2990 G2=80+((D1)/(.808))*D5
3000 G3=141.5-((D1)*(.738)*D5)
3010 G4=170.5-((D1)*(.857)*D5)
3020 A2=COS(L7)*COS(L8)/COS(D1)
3030 A1=( COS(L7)*SIN(L8)*.917463-SIN(L7)*.397821)/COS(D1)
3040 A=ATN(A1/A2)
3050 GOSUB 3390
3060 R1=A
3070 R2=R1*57.295779*24/360
3080 L1=.065709822*T1
3090 L=T*24*1.002738+6.646055+(L1-INT(L1/24)*24)
3100 LA=L-(L6*24*57.295779/360)*1.002738
3110 L=(L-INT(L/24)*24)
3120 REM: CALCULATION OF GREENWICH HOUR ANGLE (G) FROM LOCAL SIDEREAL
     TIME
3130 G=(L/24)*P5-R1
3140 IF G<P5 THEN 3170
3150 G=G-P5
3160 GOTO 3210
3170 IF G<0 THEN 3190
3180 GOTO 3210
3190 G=G+P5
3200 REM: CALCULATION OF LOCAL HOUR ANGLE (H) FROM GHA (G)
3210 H=L6-G
3220 GOSUB 4740          FOR PARALLAX CORRECTIONS
3230 REM: CALCULATION OF ELEVATION (E) OF OBJECT
3240 E3=COS(L5)*COS(H)*COS(D1)+SIN(D1)*SIN(L5)
3250 E2=SQR(1-(E3*E3))
3260 E=ATN(E3/E2)
3270 F=E
3280 IF E<0 THEN 4200
3290 IF E>16*R5 THEN 4200
3300 REM: CALCULATION OF AZIMUTH (A) OF OBJECT
3310 A2=SIN(D1)/( COS(L5)*COS(F))
3320 A2=A2-( SIN(L5)/COS(L5))*( SIN(F)/COS(F))
3330 A1=SIN(L5)*SIN(D1)+COS(L5)*COS(D1)*COS(H)
3340 A1=( SIN(H)*COS(D1))/SQR(1-A1[2)
3350 A=ATN(A1/A2)
3360 GOSUB 3390
3370 GOTO 3540
3380 REM: REMOVAL OF AMBIGUITIES INCURRED WITH ATN FUNCTION
3390 IF A=0 THEN 3410
3400 GOTO 3450
3410 IF A2<0 THEN 3430
3420 GOTO 3530
3430 A=P5/2
3440 GOTO 3530
3450 IF A>0 THEN 3510
```

(continued on page 45)

fig. 3, continued

```
3460 IF A2<0 THEN 3490
3470 A=P5+A
3480 GOTO 3530
3490 A=P5+(A-P5/2)
3500 GOTO 3530
3510 IF A2=>0 THEN 3530
3520 A=A+P5/2
3530 RETURN
3540 IF(T-I1)>(2*I)/1440 THEN 3560
3550 GOTO 3580
3560 PRINT
3570 IF WW$="YES" THEN LPRINT""
3580 Z1=INT(A*D5*10+.5)/10
3590 Z2=INT(E*D5*10+.5)/10
3600 Z3=INT(G*D5*10+.5)/10
3610 Z4=INT(D1*D5*10+.5)/10
3620 IF Z4<0 THEN 3750
3630 IF Z3<G1 THEN 3750
3640 IF Z3>G2 THEN 3660
3650 GOTO 3690
3660 IF Z3<G3 THEN 3710
3670 IF Z3>G4 THEN 3750
3680 GOTO 3730
3690 Y$="U"
3700 GOTO 3760
3710 Y$="W"
3720 GOTO 3760
3730 Y$="J"
3740 GOTO 3760
3750 Y$=" "
3760 AT$="000"
3770 BT$="00"
3780 CT$="0"
3790 BS=INT(B+.5)
3800 IF BS<10 THEN BS$=AT$+RIGHT$(STR$(BS), 1): GOTO 3840
3810 IF BS<100 THEN BS$=BT$+RIGHT$(STR$(BS), 2): GOTO 3840
3820 IF BS<1000 THEN BS$=CT$+RIGHT$(STR$(BS), 3): GOTO 3840
3830 BS$=RIGHT$(STR$(BS), 4)
3840 IF TD>0 OR TD/100=INT(TD/100) TC=TD: GOTO 3860
3850 TC=TD+2360
3860 ES=BS+TC
3870 IF ES>2400 THEN ES=ES-2400
3880 IF ES<=0 THEN ES=ES+2400
3890 IF ES<10 THEN ES$=AT$+RIGHT$(STR$(ES), 1): GOTO 3930
3900 IF ES<100 THEN ES$=BT$+RIGHT$(STR$(ES), 2): GOTO 3930
3910 IF ES<1000 THEN ES$=CT$+RIGHT$(STR$(ES), 3): GOTO 3930
3920 ES$=RIGHT$(STR$(ES), 4)
3930 IF LA<0 THEN LA=LA+24
3940 IF LA>24 THEN LA=LA-24
3950 LB=100*INT(LA)
3960 LC=60*(LA-INT(LA))
3970 IF LC-INT(LC)=>0.5 LC=INT(LC)+1 ELSE LC=INT(LC)
3980 IF LC=>60 LC=0: LB=LB+100
3990 LD=LB+LC
4000 IF LD>2400 THEN LD=LD-2400
4010 LB$=STR$(LD)
4020 IF LD<10 THEN LB$=AT$+RIGHT$(LB$, 1): GOTO 4060
4030 IF LD<100 THEN LB$=BT$+RIGHT$(LB$, 2): GOTO 4060
4040 IF LD<1000 THEN LB$=CT$+RIGHT$(LB$, 3): GOTO 4060
4050 LB$=RIGHT$(LB$, 4)
4060 Z1$="###.#"
4070 Z2$="##.#"
4080 Z3$="###.#"
4090 Z4$="+##.#"
4100 PRINT USING"%  %";BS$;
4110 PRINT TAB(7)USING Z3$; Z3;
4120 PRINT TAB(16)USING Z4$; Z4;
4130 PRINT Y$;
4140 PRINT TAB(27)LB$;
4150 PRINT TAB(37)USING"%  %";ES$;
4160 PRINT TAB(45)USING Z1$; Z1;
4170 PRINT TAB(55)USING Z2$; Z2
4180 IF WW$="YES" THEN LPRINT USING"%  %";BS$;: LPRINT TAB(7)USING
     Z3$; Z3;: LPRINT TAB(16)USING Z4$; Z4;: LPRINT Y$;: LPRINT TAB(27)LB$;
     : LPRINT TAB(37)USING"%  %";ES$;: LPRINT TAB(45)USING Z1$; Z1;:
     LPRINT TAB(55)USING Z2$; Z2
4190 I1=T
4200 B=B+I
4210 Z=(B-INT(B/100)*100)-60
4220 IF Z<0 THEN 2730
4230 B=INT(B/100)*100+100+Z
4240 GOTO 2730
4250 RX=R2
4260 IF RX<0 THEN RX=RX+24
4270 IF RX>24 THEN RX=RX-24
4280 RA=100*INT(RX)
4290 RB=60*(RX-INT(RX))
4300 IF RB-INT(RB)=>0.5 RB=INT(RB)+1 ELSE RB=INT(RB)
4310 IF RB=60 RB=0: RA=RA+100
```

(continued on page 47)

TRS-80 to select days and months for 1978 and 1983. Why did I select those particular years? 1983 was the current year; 1978 is 5 years before. In addition, copies of the *Nautical Almanac* for each of those years were easily available. I printed out the moon coordinates to three decimal places for the selected dates. Then I modified an existing statistical program so that I could efficiently use the TRS-80 to calculate the average, or arithmetical means, standard deviations, and ranges of errors.

Average errors of less than 0.01 degree were obtained with maximum errors under 0.2 degree in sample sizes approaching 100. This inherent accuracy of the basic program equations should be adequate for many Amateur EME stations.

The above analysis is for geocentric GHA and declination data with the *Nautical Almanac* used as an accurate standard for comparison. I could find no comparable standard for azimuth and elevation error analysis. However, at least one checkpoint each day can be used to gain some confidence in the AZ/EL calculations. When the moon's GHA equals your longitude, the azimuth should be 180 degrees. At that same time the elevation should be equal to 90 degrees minus your latitude (+north/−south) plus the declination (+north/−south). Checking these points for the days used in the preceding analysis, it was found that the azimuth differences at 180 degrees were less than 0.011 degrees and elevation under 0.001 degrees, maximum. Average errors were 0.00037 and 0.00013, respectively. (These are for geocentric values.)

Having gained confidence in the ability of the fundamental equations to provide accurate results, the parallax corrections which had been bypassed for the above analyses were reinstated to provide output values referred to the surface rather than the center of the earth. These are given in lines 4750 through 4850 and were based on information found in reference 3. Again, the lack of an accurate standard precludes a statistical error analysis but a straight comparison of GHA, declination, azimuth and elevation values before and after the parallax corrections may be useful.

When this was done, all corrections were less than 1.0 degree. Those for azimuth were "in the noise level," i.e., less than 0.005 degree. At moonset and moonrise, the elevation correction factors were very close to 1 degree and at azimuths near 180-degrees about 0.6 degree. (A detailed error analysis is available from *ham radio*, Greenville, NH 03048. Send an SASE with two first-class stamps attached. Request W2WD Moon Coordinate Error Analysis.)

optimizing the program

The program as presented is arranged for universal use. For frequent use at one location, it should be streamlined to eliminate the time-consuming task of keying in the repetitive portions of the input data. To do this, the following changes should be made:

- At **line 1450**, remove the apostrophe ('). This has the effect of activating this GOTO 1730, which bypasses the repetitive input data requirements.
- At **line 1730**, remove the apostrophe (') and activate this line, replacing my standard input data with your standard input data:

W$	= "your call letters" (use quotation marks)
L5	= your latitude, degrees [negative (−) for south latitude]
U5	= your latitude, minutes
L6	= your longitude, degrees [negative (−) for east longitude]
U6	= your longitude, minutes
TL$	= "your time zone" (use quotation marks)
TD	= time differential from GMT to your zone [negative (−) for zones west of GMT]

When these changes have been made, the program, when called up, will immediately go to the question WHAT IS THE DESIRED PRINTING INCREMENT IN MINUTES (1-60)? Thus, you have avoided the logo, the need to input your call letters, local time, time differential, latitude, longitude. This saves a lot of data entry time. I use two tailored programs . . . one set for Eastern Standard Time, for which I use the filespec "MOONEST" and the other "MOONEDT" for Eastern Daylight Savings Time. If you never use the low-elevation mode, this section can be bypassed. If you do not have a printer, then that question can likewise be skipped. With some knowledge of BASIC programming, making changes to suit your individual requirements are not difficult.

The program was rewritten for the level II dialect of the Model I TRS-80 from versions presented in reference 4 by Lance Collister (WA1JXN) and Jay Liebmann (K5JL). Enhancements have been added in the form of readouts for Local Sidereal Time, Right-Ascension of the moon and path loss variations. Provisions have also been made for the direct inputting of time zones and time differentials. Keyboard input statements have been error-trapped to reduce the chance for "cockpit" errors.

For those wishing to avoid the task of manually keying in the 10,000-byte program listing (see **fig. 3**), a 500-baud, level II BASIC cassette tape is available from the author for $15.00 including postage. For Disk BASIC the taped program may be "cloaded" into memory after first shutting off the interrupts with the CMD"T" command. From memory it may then be "saved" onto a disk for future use. Two versions are on the tape. The first is the version as presented in this article. It was stretched out to make it easier to list and modify to your own requirements. The second

fig. 3. continued

```
4320 RC=RA+RB
4330 RY$=STR$(RC)
4340 IF RC<10 THEN RY$=AT$+RIGHT$(RY$, 1)
4350 IF RC<100 THEN RY$=BT$+RIGHT$(RY$, 2)
4360 IF RC<1000 THEN RY$=CT$+RIGHT$(RY$, 3)
4370 RY$=RIGHT$(RY$, 4)
4380 PRINT
4390 IF WW$="YES" THEN LPRINT""
4400 PRINT"R.A. OF MOON = ";RY$; "      PATH-LOSS INCREASE  +";DB
     " DB"
4410 IF WW$="YES" THEN LPRINT "R.A. OF MOON = ";RY$;"      PATH-LOSS
     INCREASE  +";DB;" DB"
4420 PRINT
4430 NEXT N
4440 END
4450 REM: CALCULATE DISTANCE TO THE MOON
4460 DD#=JD#-2444238.5
4470 AA=0.98564733
4480 ED=-3.76286
4490 MS=(AA*DD#)+ED
4500 IF MS<0 THEN MS=MS+360: GOTO 4500
4510 IF MS>360 THEN MS=MS-360: GOTO 4510
4520 AE=0.1858*SIN(MS*0.0174533)
4530 AF=0.37$SIN(MS*0.0174533)
4540 LS=(AA*DD#)+(1.9157417*SIN(0.0174533*((AA*DD#)+(-3.76286))))
     +278.833540
4550 IF LS<0 THEN LS=LS+360: GOTO 4550
4560 IF LS>360 THEN LS=LS-360: GOTO 4560
4570 LL=(13.1763966*DD#) + 64.975464
4580 IF LL <0 THEN LL=LL+360: GOTO 4580
4590 IF LL >360 THEN LL=LL-360: GOTO 4590
4600 CC=LL-LS
4610 MM=LL-(0.1114041*DD#)-349.383063
4620 IF MM<0 THEN MM=MM+360: GOTO 4620
4630 IF MM >360 THEN MM=MM-360: GOTO 4630
4640 EV=1.2739*SIN(((2*CC)-(MM))*0.0174533)
4650 MN=MM+EV-AE-AF
4660 EC=6.2886*SIN(MN*0.0174533)
4670 MD=383242.41/(1+0.054900*COS((MN+EC)*0.017453292))
4680 REM: CONVERT DISTANCE VARIATION TO PATH-LOSS CHANGE (DB)
4690 DB=MD/356334
4700 DB=40*LOG(DB)/LOG(10)
4710 DB=INT(DB*10+0.5)/10
4720 RETURN
4730 END
4740 REM; CORRECTIONS FOR PARALLAX
4750 H1=H
4760 R#=MD/6378.16
4770 U=ATN(0.996647*TAN(L5))
4780 P1=0.996647*SIN(U)
4790 P2=COS(U)
4800 HC=ATN((P2*SIN(H))/(R#*COS(D1)-P2*COS(H)))
4810 H=H+HC
4820 G=L6-H
4830 D1=ATN((COS(H))*((R#*SIN(D1))-P1)/((R#*COS(D1)*COS(H1))-P2))
4840 RETURN
```

is a customized version which will be set to your station parameters if you include the necessary information. I will need your station call letters, your latitude, your longitude, your local time zone, and the difference between your time and GMT. Also include whether you want the low-elevation and printing routines bypassed.

The two programs are also available on a diskette ($18.00) formatted to run on TRSDOS compatible operating systems. It has been checked on single-density TRSDOS 2.3, NEWDOS 2.1, NEWDOS-80 1.0, DOSPLUS 3.3 and LDOS 5.0. A double-density version will run on NEWDOS-80 2.0 or DBLDOS 4.23 and perhaps others using a Percom doubler with a Model I machine. Be sure to specify single or double-density and include the same information as specified in the paragraph above.

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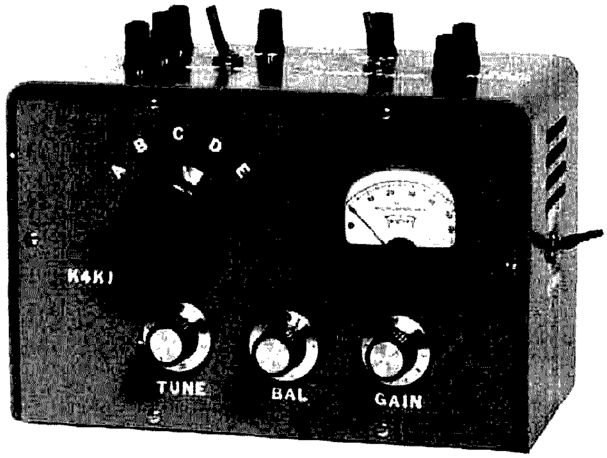
ham radio

Use this device
to determine Yagi
or other antenna patterns



a sensitive field strength meter

While attempting to make some meaningful relative field strength measurements on an antenna system some months ago, I found my years-old field strength meter to be quite inadequate. Its lack of sensitivity meant that measurements had to be made close to the antenna, where the induction field introduced error in the radiation field I was actually interested in. Its non-linearity was also a problem.

I decided to design a new unit that would include a stable, sensitive linear amplifier, provisions for remote monitoring, and a meter calibrated in decibels. The final results proved so satisfactory that it seemed other Amateurs would be interested in this project.

Figure 1 shows the schematic. The two-pole, five-position switch, coils and 365 pF variable capacitor cover a range from 1.5 to 30 MHz. (The combination of parts came from an old low-power transmitter.)

Almost all handbooks show coil-capacitor combinations that can be used for field strength meters, so Amateurs can build the kind of unit that best suits their needs. In addition, they can incorporate any kind of pickup coupling system for the pickup antenna of their choice. My own measurements were sufficiently satisfactory using a short antenna only a few feet long connected directly to the ANT binding post. When a longer antenna for pickup was desired, it was connected to CAP, binding post C, where it connected to C2, a small 50 pF variable capacitor.

The amplifier uses a couple of Darlington NPN transistors whose high beta, 5000, provides high sensitivity with S1 used as the amplifier ON/OFF switch. Switch S2 in the left position allows the output of the 1N34 diode to be fed directly into the 50 μA meter (M) for direct reading. When S2 is in the right position, the amplifier is switched into the circuit. Switch S3 is for LOCAL or REMOTE monitoring. At full GAIN setting the input signal is adjusted to give a full-scale reading of fifty microamperes on the meter. Then with the amplifier switched out of the circuit, the meter reading drops down to about half a microampere.

A 2.5 mH RF choke and capacitors C3, C4, and C5 effectively keep RF out of the amplifier circuit.

two balance adjustments are required

Because of the high sensitivity of the amplifier circuit, it's best to have two balancing controls, R7 (BAL, front panel adjustable) and R8 (an internal screwdriver-adjustable potentiometer). When initially setting R8, adjust it in conjunction with R7. Doing this allows the GAIN control to be varied from zero to full gain with the meter indication remaining at zero.

Because the amplifier circuit is basically a balanced bridge type, a zero-center meter could have been used. But none was available, so R3, an 82-ohm offset resistor, was added. Its function is to prevent the meter from swinging too hard to the left below zero, which might occur if the panel balance control were inadvertently turned fully counterclockwise with the amplifier gain turned fully on. The best safe operating procedure to employ is always to turn the gain con-

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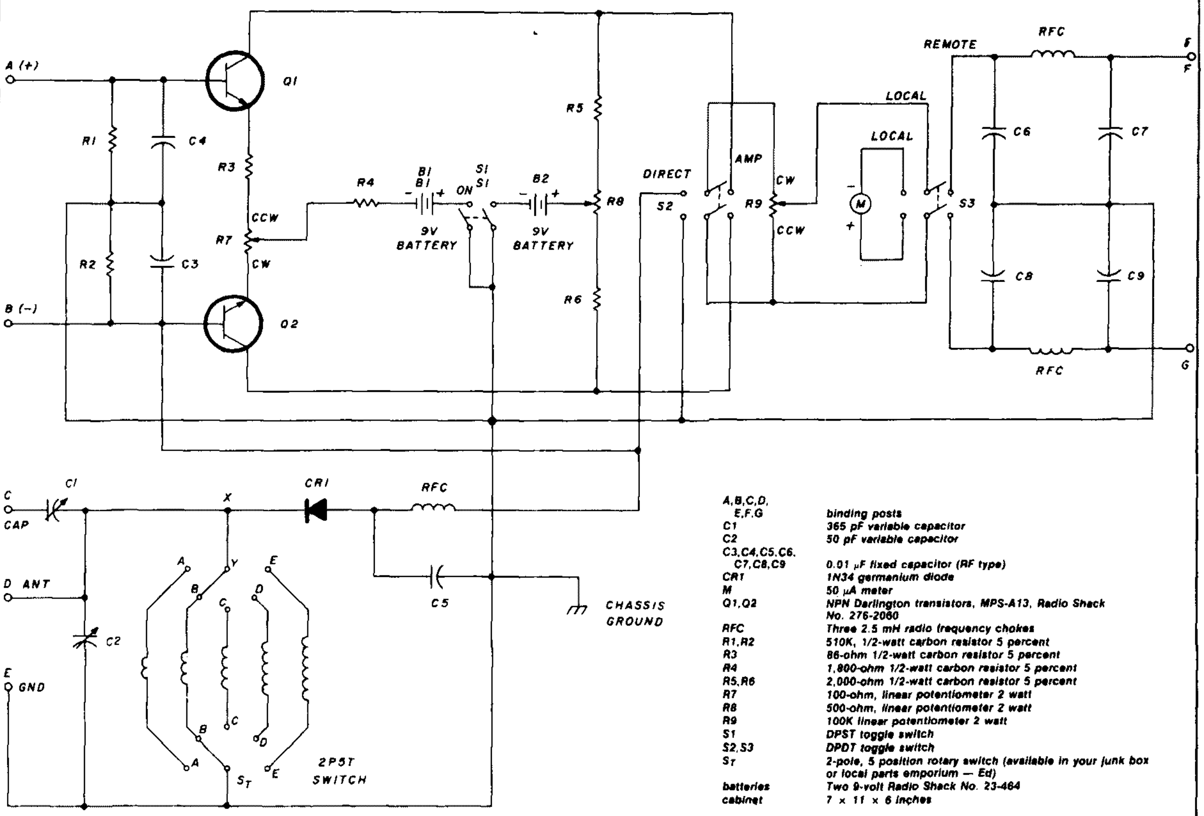


A,B,C,D, E,F,G	binding posts
C1	365 pF variable capacitor
C2	50 pF variable capacitor
C3,C4,C5,C6, C7,C8,C9	0.01 μF fixed capacitor (RF type)
CR1	1N34 germanium diode
M	50 μA meter
Q1,Q2	NPN Darlington transistors, MPS-A13, Radio Shack No. 276-2060
RFC	Three 2.5 mH radio frequency chokes
R1,R2	510K, 1/2-watt carbon resistor 5 percent
R3	86-ohm 1/2-watt carbon resistor 5 percent
R4	1,800-ohm 1/2-watt carbon resistor 5 percent
R5,R6	2,000-ohm 1/2-watt carbon resistor 5 percent
R7	100-ohm, linear potentiometer 2 watt
R8	500-ohm, linear potentiometer 2 watt
R9	100K linear potentiometer 2 watt
S1	DPST toggle switch
S2,S3	DPDT toggle switch
S_T	2-pole, 5 position rotary switch (available in your junk box or local parts emporium — Ed)
batteries	Two 9-volt Radio Shack No. 23-464
cabinet	7 × 11 × 6 inches

fig. 1. Field strength meter schematic.

trol down to zero before turning the power switch ON, and then initially balance the circuit while slowly advancing the gain control. This is a common precaution used in all sensitive bridge circuits.

additional features

The addition of binding posts A and B allows the amplifier to be used by itself as a high impedance, high gain amplifier, where binding posts A and E are used for a + input and ground.

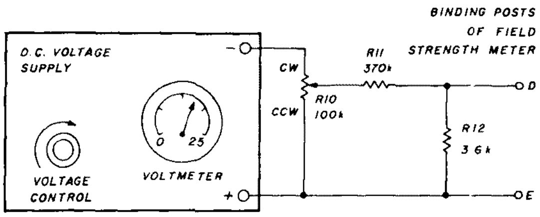
Switch S3, a remote metering circuit in my original field strength meter, was also incorporated in the new, more sensitive unit. However, it was found necessary to add the RF filter circuit to help eliminate RF pickup from the remote metering wires to the unit. This remote metering feature is not really necessary, but is handy if you want to elevate the field strength meter a few feet when making beam antenna measurements. Care should be taken that the remote meter wires are both shielded and properly grounded, because it doesn't take much RF getting back into the field strength meter to upset the initial balance.

Although the meter readings by themselves are a good indication of relative field strength, it was thought desirable to provide a theoretical calibration curve relating meter deflection to dB change, and then to do an actual calibration using the test circuit shown in fig. 2.

calibrating is easy

The theoretical dB calibration curve can be best understood if, for example, we use full scale on the meter 50 μA as 0 dB with a given input signal. Now if the input signal were reduced so that the meter reads 30 μA, then the dB drop in signal level is dB = 20 $\log_{10}$ (30/50) = 4.44 dB. This calculation is for a perfect linear system; and by using this mathematical procedure it is easy to calculate and develop the theoretical dB curve of fig.3 for different values of meter current.

This curve is very easy to use. For example, with the field strength meter set up away from your station, if you rotated your beam and saw that the maximum meter reading went over 50μA, just cut back on the GAIN control on the field strength meter, so that the maximum reading is either 50 μA or somewhat lower. The maximum reading does not have to be exactly 50 μA, as shown by the following example.



item	description
PS	0-25 volt DC power supply (see text)
R10	100 kilohm 2-watt potentiometer
R11	370 kilohm, 1/2-watt carbon resistor
R12	3.6 kilohm, 1/2-watt carbon resistor

fig. 2. Calibration test unit.

If the maximum reading were 45 μA and the minimum reading 10 μA, then from **fig. 3**, 45 μA is equal to -0.915 dB, and 10 μA is equal to -13.98 dB and the numerical difference is -13.065 dB, which can be rounded off to -13.1 dB.

The reason that it is not necessary to have the microammeter set exactly at 50 μA when reading the maximum signal is that decibels can be added or subtracted from each other anywhere along the curve to give the correct dB difference.

It should be pointed out, however, that better accuracy is obtained if the maximum reading is set initially close to 50 μA. This is because, as in most metering systems, the percentage accuracy obtainable is less at the lower end of a scale than at the top. Also, the curve shown was not *extended* below 5 *m*A.

I happened to have a small variable 0-25 VDC power supply with a reasonably accurate built-in voltmeter. Almost any kind of variable supply with a voltmeter can be used, as the power supply voltage is factored into the calibration procedure to be described. The attenuator network is used to reduce the power supply voltage so that full scale on the microammeter could readily be obtained at full gain of the amplifier, with the power supply set at 25 volts.

Before beginning the calibration procedure, it is first necessary to isolate the resonant circuits from the input terminals; otherwise the coils would provide a DC short between the RF input binding posts D and E used for the calibration input voltage.This can be accomplished by temporarily opening the circuit between points X and Y of **fig. 1**. An alternate method is just to slip a piece of paper between the coil contact on the turret and the arm of the rotary switch.

The test unit is then connected to the input binding posts D and E, as shown, and R10 rotated to the counterclockwise, or zero-output position. The amplifier is now turned on, properly balanced to zero, and the GAIN control set to its maximum position. The balance should again be checked, and it should still be at zero. The power supply is turned on, and the voltage adjusted to read 25 volts. R10 is rotated in a clockwise direction until the field strength microammeter indicates a full scale reading of 50 μA. R10 should not be adjusted further, since its setting now has established the zero dB level.

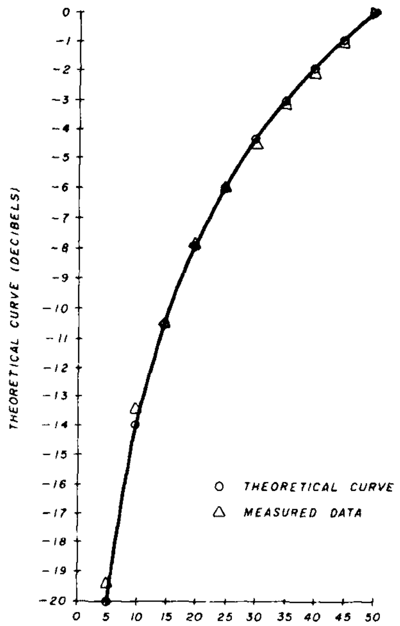
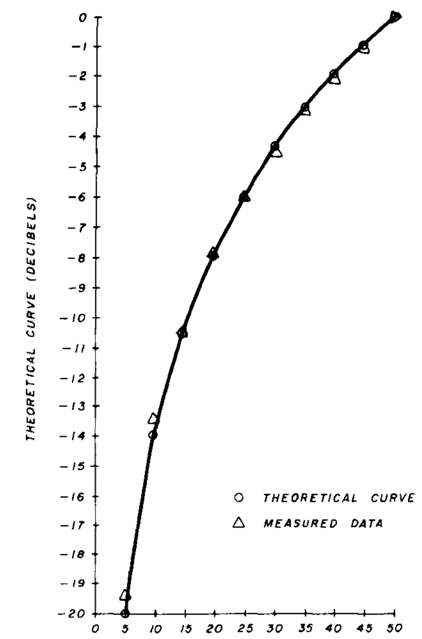


fig. 3. Measured data showing close correlation to theoretical values indicates good amplifier linearity.

The power supply voltage is now reduced until the microammeter reads 45 μA. The power supply voltage is now read and recorded. This procedure is again repeated for 40 μA, again reading and recording the power supply voltage. Similar successive steps, each 5 μA lower are done, until the final microampere reading of 5 μA is made and data recorded. The actual recorded data is indicated in **fig. 3**.

The actual calculated dB for a measured reading of voltage is done as follows: for example, at 30 μA, the power supply voltage was found to be 15.2 volts. The actual calculated dB for a measured reading of, for example, 30 μA, and a recorded power supply voltage of 15.2 volts is equal to $20 \log_{10} (15.2/25) = -4.32$ dB. This is quite close to the theoretical value of -4.44 dB previously shown. The resulting data and curve is shown in **figs. 3** and **4**. Good correlation between theoretical and observed data shows that the amplifier had a good linear characteristic. Although the amplifier was calibrated using DC voltage, the results obtained showed the unit to be very useful as a device for measuring relative dB levels. And that is what most Amateurs are really interested in. The measure of absolute field strength normally requires extremely good



MICROAMPERES	THEORETICAL dB	TEST POWER SUPPLY VOLTAGE	CALCULATED MEASURED dB
50	0	25	0
45	-0.915	22.4	-0.954
40	-1.94	19.7	-2.07
35	-3.10	17.6	-3.05
30	-4.44	15.2	-4.32
25	-6.02	12.5	-6.02
20	-7.96	10.1	-7.87
15	-10.46	7.6	-10.34
10	-13.98	5.3	-13.47
5	-20	2.6	-19.66
0	∝	——	——

fig. 4. Measured and theoretical data.

commercial equipment and is not generally used in Amateur-type measurements.

Although the calibration curve was shown for the full maximum gain of the amplifier, several similar curves were run at lower gains, down to 20 percent of maximum gain, and no noticeable variations were encountered. This indicates that the curve shown can be used between the limits of full gain down to 20 percent of full gain. My own experience is that if your signal is very strong, just reduce the size of your pickup antenna or move your field strength meter further away from your transmitting antenna. Actually, the further you are away from your transmitting antenna, the better patterns you will obtain.

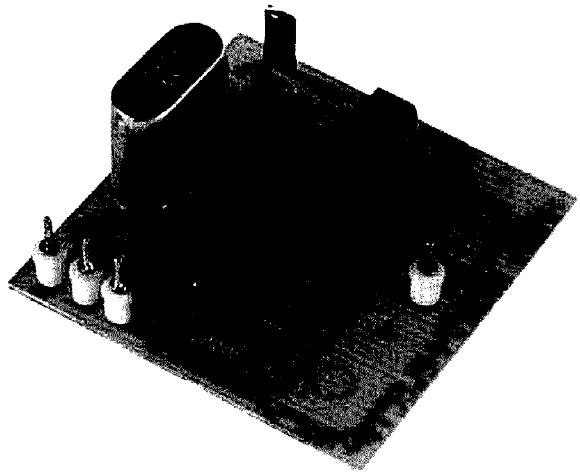
field testing

After experimentation at my own station, the unit was taken to the home of Russell Forsyth, K4YS, who had recently put up a large 20-meter beam. The azimuthal pattern in dB was readily obtained, with a front-to-back ratio measured at 14 dB, corresponding to a power ratio of a little over 25:1, which seemed reasonable for the beam he used. A rather interesting anomaly was observed, in that at a certain direction when the beam was rotated, the field strength meter showed a marked irregular intermittent variation. Shaking the tower seemed to aggravate the condition, indicating that perhaps one of the elements was loose, or perhaps the mast was not perfectly vertical, and at certain positions, perhaps loose bolts or something else was causing the problem. At the moment, we still do not know what the problem is, but it does show that the field strength meter can also be used as a diagnostic tool in problems of this type.

Other tests near my home, where we have a high-voltage line nearby, showed that there was some sort of pickup near the power poles on rainy days, particularly when corona could be heard. The noise was tunable and verified as line noise interference by using a small transformer and earphones connected to the REMOTE binding posts instead of a meter. Apparently the lightning protection system, which incorporates a ground at each pole and a overhead ground line, was being shock-excited by the corona discharge and causing those loop circuits to have induced noise oscillations. (At least that's what it seemed to me.) Perhaps later I can use the meter to track down excessive line noise that occasionally is quite high on my Yaesu FT-101B.

There were no problems in the design or construction of the basic unit and its modifications. Though the field strength meter, as shown in the photograph, is quite large, it could certainly be reduced in size. I'm hoping that its use will provide Amateurs with a more sensitive method of making relative antenna measurements, so essential to good antenna experimentation. I'll be glad to answer any questions or comments you might have. Just send an SASE to the author at the address indicated.

ham radio



Modern design approach
cuts cost, reduces size,
and eliminates drift,
tuning, tweaking,
and passband loss

an integrated circuit low-pass filter

Operational amplifiers, developed and popularized in the late 1960s, have all but eliminated audio amplifiers designed using discrete transistors, resistors, and capacitors. Amplifiers built with discrete components are usually designed for special applications. In an analogous manner, audio frequency filters have evolved. First came the LC filters. In widespread use since before 1920, they are characterized by large physical size and weight, complexity of design procedure, and critical component values. Next came the active filters. Active filters were first implemented using vacuum tubes, then discrete transistors, and today using integrated circuits — typically quad op amps. These circuits did away with bulky inductors and complicated design procedures. Any adjustments needed on an active filter could be accomplished by "tweaking a pot" or two. As manufacturers of integrated circuits saw large numbers of op amps being used to manufacture the same type of active filters again and again, they saw the opportunity to profit from the manufacture of dedicated filtering devices. The use of one of these integrated circuit filters is the subject of this article.

The device discussed in this article is the S3528 programmable low-pass filter, manufactured by American Microcircuits, Inc., 3800 Homestead Road, Santa Clara, California 95051. Contained within one eighteen-pin dual inline package (DIP) is a complete seventh-order elliptical low-pass filter; its passband ripple is less than 0.1 dB, and its stopband attenuation is greater than 51 dB for frequencies greater than 1.3 f_c (cutoff frequency).

In addition, this IC contains two uncommitted operational amplifiers that can be used to provide additional filtering and gain. The device is also progammable for cutoff frequency. This means that the cutoff frequency is not fixed, but can be changed by providing switch or logic level inputs to the device. The S3528 also has a built-in oscillator for use with an external crystal,

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Block Diagram **Pin Configuration**

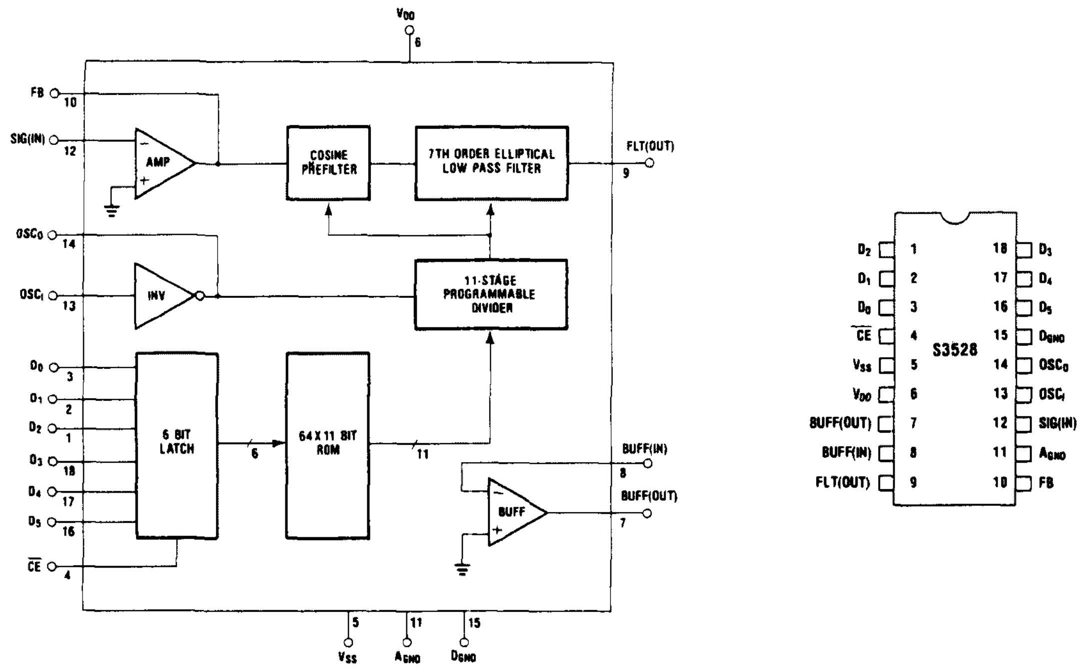

fig. 1. S3528 programmable low-pass filter. *(Courtesy American Microsystems, Inc., 1984.)*

such as the low-cost 3.58 MHz TV crystal. This feature allows very stable performance over a wide range of temperature and voltage. If desired, an external oscillator may be used. The frequency range of operation of this low-pass filter is 40 Hz to 20 kHz with the TV crystal, or 10 Hz to 20 kHz with an external oscillator. A block diagram and pin layout are shown in **fig. 1**.

The six data inputs, DØ-D5, are tied to either +5 VDC or GND to program the cutoff frequency into the device according to the coding shown in **table 1**. These data inputs, latched by an input buffer, are used to preset the internal divider to give the proper frequency to the switched capacitor filter.

In our demonstration circuit, a 6-position jumper "patch panel" provides the frequency selection. If you connect the DØ through D5 inputs to the buffered data bus of a microcomputer, the cutoff frequency can be varied by simply outputting the proper code per **table 1**.

switched capacitor filters

This integrated circuit filter uses CMOS op amps, CMOS transmission gates (switches), and CMOS capacitors to synthesize the required filtering function. The technique used is that of a switched capacitor filter. The following discussion will explain the functioning of a simple SCF circuit.

Figure 2 illustrates a simple RC active first order low-pass filter and **fig. 3** an SCF version of the same circuit. If the switch in **fig. 3** is in position "A," the voltage on capacitor C_r becomes equal to the input voltage. If the switch is then changed to position "B," the charge on C_r can be considered to have been transferred to capacitor C. The basic function illustrated being that of integration or low-pass filtering.

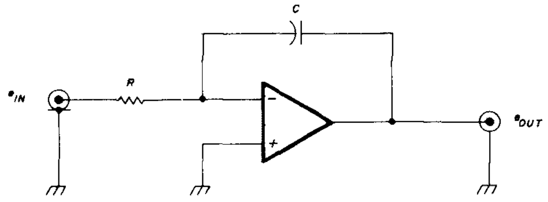

fig. 2. Simple RC-active low-pass filter.

The value of the simulated resistor (C_r and the SPDT switch) depends only on the capacitance of C_r and the switching frequency. Since the ratio of capacitor sizes is proportional to their surface areas in an integrated circuit, the ratio remains stable over a wide temperature and voltage range. If the switching frequency is crystal controlled, or otherwise stabilized, a very frequency-stable filter results.

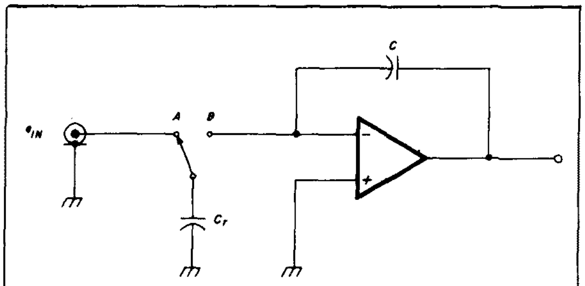

fig. 3. Switched capacitor equivalent to *fig. 2*.

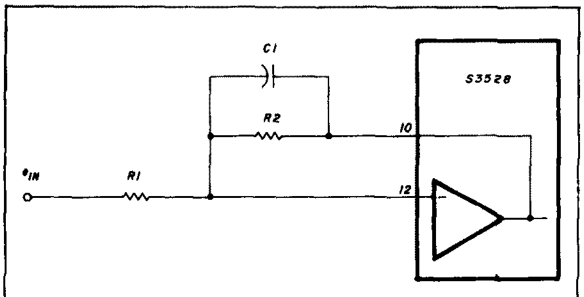

fig. 4. Input anti-alias low-pass filter.

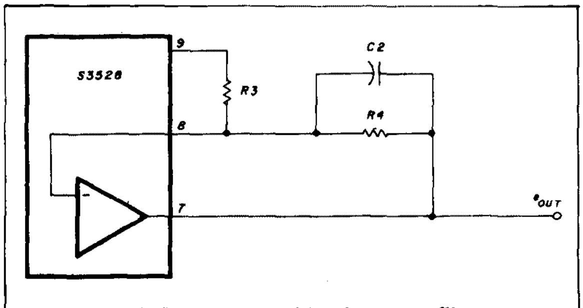

fig. 5. Output smoothing low-pass filter.

one problem with SCFs

The S3528 filter uses a switching frequency approximately 80 times the progammed cutoff frequency. As long as the input signals are far removed, no greater than 10 percent of the switching frequency, there are no problems. If, however, some portion of the input signal contains energy at the switching frequency, a phenomenon known as "aliasing" will occur. An alias is a spurious signal that appears at the output, indistinguishable from the desired low-frequency signal.

One way to visualize what is happening is to think of the stroboscopic effect observed in a Western movie when the frame flicker rate is approximately equal to the rotational speed of the spoked wagon or stage coach wheels: the wheels may appear to be stationary, or even going backward. This is an example of visual aliasing.

To prevent this from happening in our circuit, we apply a simple active RC low-pass filter to the input. We use the input op amp as shown in **fig. 4**.

To properly size R1, R2, and C1 the following method is used:

Determine the required amplification factor of the filter. For ease of application, limit gains to between 1 and 10. Assume R1 = 10 kilohms. This establishes the input impedance of this filter. R2 = n·R1 where n is the required gain.

Determine the desired cutoff frequency. Multiply this frequency by 2 and apply this to the following formula:

$$C = \frac{1}{6.28 \cdot f \cdot R2} \qquad (1)$$

For example:

Gain = 3.5

Cutoff frequency = 100 Hz

R2 = 35 kilohms

table 1. Frequency vs. input coding for S3528.

NOMINAL CUTOFF FREQ. (HZ)	INPUT CODE BINARY (HEX) D5-D0	DIVIDER RATIO	ACTUAL CUTOFF FREQ. (HZ)
40	000000 (00)	2048	44
100	000001 (01)	895	100
200	000010 (02)	447	200
250	001011 (0B)	358	250
300	000011 (03)	298	300
400	000100 (04)	224	399
475	001010 (0A)	188	476
500	000101 (05)	179	500
600	000110 (06)	149	601
700	000111 (07)	128	699
800	001000 (08)	112	799
900	001001 (09)	99	904
1000	001100 (0C)	90	994
1000	010000 (10)	89	1005
1030	001101 (0D)	87	1028
1050	001110 (BE)	85	1053
1100	010001 (11)	81	1105
1150	001111 (0F)	78	1149
1200	010010 (12)	74	1209
1300	010011 (13)	69	1297
1400	010100 (14)	64	1398
1470	011010 (1A)	61	1467
1500	010101 (15)	60	1491
1540	011011 (1B)	58	1542
1600	010110 (16)	56	1598
1700	010111 (17)	53	1688
1720	011100 (1C)	52	1721
1800	011000 (1B)	50	1790
1900	011001 (19)	47	1904
1950	011101 (1D)	46	1945
2000	100000 (20)	45	1989
2030	011110 (1E)	44	2034
2100	100001 (21)	43	2081
2200	100010 (22)	41	2183
2240	011111 (1F)	40	2237
2300	100011 (23)	39	2295
2350	101010 (2A)	38	2350
2400	100100 (24)	37	2418
2500	100101 (25)	36	2686
2560	101011 (2B)	35	2557
2600	100110 (26)	34	2632
2700	100111 (27)	33	2711
2800	101000 (28)	32	2797
2900	101001 (29)	31	2887
3000	110000 (30)	30	2983
3100	110001 (31)	29	3086
3200	110010 (32)	28	3196
3300	110011 (33)	27	3314
3400	110100 (34)	26	3442
3600	110110 (36)	25	3579
3700	110111 (37)	24	3728
3900	111001 (39)	23	3891
4000	101100 (2C)	22	4067
4470	101101 (2D)	20	4474
5000	101110 (2E)	18	4971
5600	101111 (2F)	16	5593
6000	110101 (35)	15	5965
6400	111000 (38)	14	6392
7500	111010 (3A)	12	7457
9000	111011 (3B)	10	8949
10000	111100 (3C)	9	9943
15000	111101 (3D)	6	14915
18000	111110 (3E)	5	17897
22000	111111 (3F)	4	22372

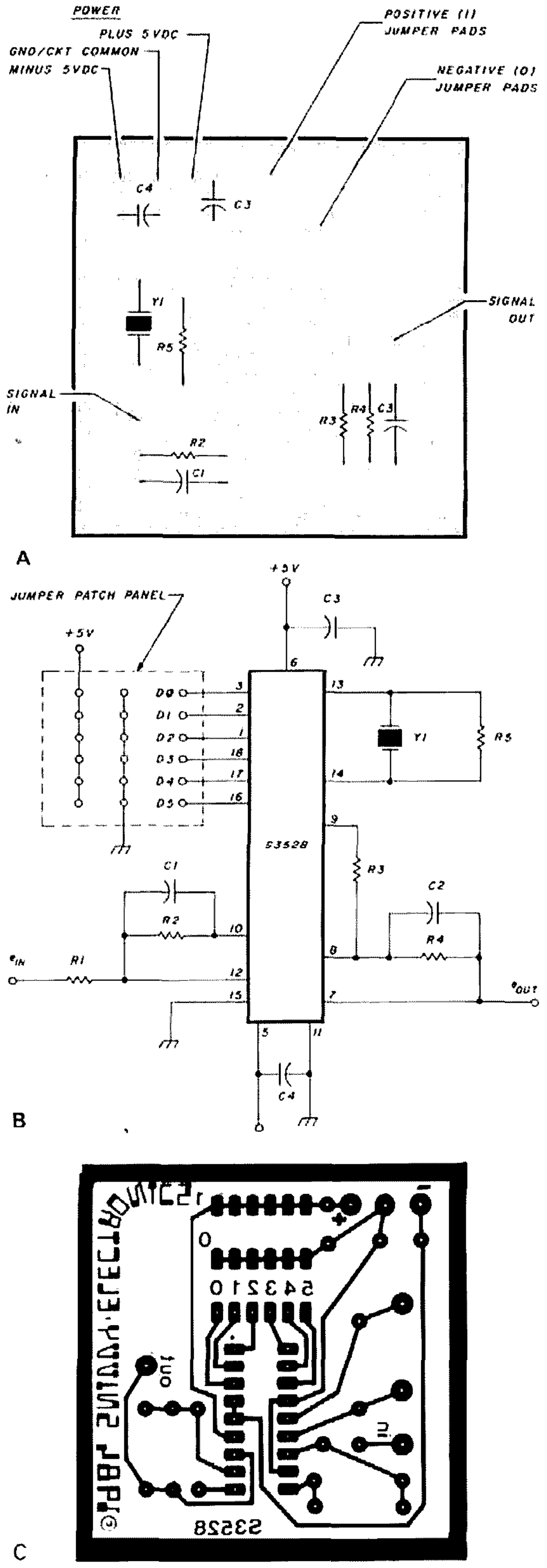

fig. 6. (*A*) shows parts layout; (*B*), wiring diagram; (*C*), 1:1 negative PC layout.

item	description
C1,C2	determined by the desired frequency, pollster, 35 VDC
C3,C4	0.01 μF, disc ceramic, 35 VDC
IC1	S3528 AMI, Inc.
R1,R2,R3,R4	10 kilohm, 1/4-watt, carbon film resistors
R5	10 megohm, 1/4-watt, carbon film resistor
socket	18 pin, solder tab type
Y1	3.58 MHz "color burst" crystal, HC-6-U holder
Either printed circuit card or perforated wiring board	
Miscellaneous: solder, 12-inch tinned, bare jumper wire	

$$C = \frac{1}{6.28 \cdot 200 \cdot 35000} = 0.02\ \mu F$$

The values of R and C calculated are not critical. You can use the nearest available device without significant performance degradation.

The output of the S3528, pin 9, should not be used directly to drive a load of less than 10 kilohms. To buffer the output and to provide smoothing of the output signal, the op amp at pins 6 and 7 can be configured as a low-pass filter. Refer to **fig. 5** for details.

To properly size R3, R4, and C2, use the following guidelines and example:

gain of output op amp/filter = 1
R3 = R4 = 10 kilohms
output low-pass filter corner frequency
= 2·elliptic filter cutoff frequency

Calculate:

$$C = \frac{1}{6.28 \cdot R3 \cdot 2 \cdot f} = 0.079\ \mu F$$

(0.1 μF is close enough)

For applications requiring a variable cutoff frequency over a 10:1 (or less) range, the input and output networks should be sized for the highest desired frequency. For applications over a wider range, external filtering or using switched filtering components may be required, depending on the nature of the signals being processed.

As with any new components, it is wise to experiment with the S3528 on a solderless breadboard or a prototype evaluation board prior to incorporating it into a new design.

Readers who wish to etch their own PC cards should refer to **fig. 6** for a full-size copy of the negative artwork and components layout. Printed circuit cards and kits are available from the author; see page 63 for details.

circuit applications

The circuit and PC card have been used in numerous applications around the ham shack and laboratory. As a low-pass filter for preprocessing the signals from the station microphone, it gives good performance while having extremely fast rolloff above the cutoff frequen-

cy. As an audio output processor from the station receiver, it is capable of narrowing the effective width of the receiver's internal crystal filters without degrading the shape factor of the crystal filters.

In low-frequency signal processing, it can completely eliminate 60 Hz audio hum from low-level signals. It could be used in a modem or terminal unit for this purpose. Similarly, it can be used in biomedical or biofeedback circuits to eliminate hum and higher frequency products while extracting EEG or ECG signals.

conclusion

The device described in this article is only one of many switched capacitor filter integrated circuits available for use as general-purpose SCF building blocks for producing literally any complex filtering function. Designers of commercial and military equipment are making good use of these devices for lower cost, higher reliability, reduced size, reduced weight, and ease of design.

Judgment must be used when using this filter with pulse or digital circuits. As with any high Q filter, ringing and pulse shape distortion may occur. Again, experiment with the filter before committing to a new design.

The following are available from Snivvy Electronics, 45 Salem Lane, Little Silver, N. J. 07739.

Printed circuit card, glass epoxy	**$ 6.50**
S3528 integrated circuit	**$12.50**
Complete kit of all parts including capacitors C1, C2 for 100 Hz, 1000 Hz, 2000 Hz, 2500 Hz	**$22.50**

N. J. residents please add 6 percent N. J. Sales Tax. All sales include a copy of the S3528 applications brief and complete circuit data.

ham radio

the reduced-size, full-performance, corner-fed delta loop

For broadband DX performance, try this one on for size

The full wave loop, whether in delta or quad form, has enjoyed popularity over many years. It forms the basis for the well-known and very effective 2-element quad antenna, the history of which can be traced back over 30 years.

A single loop in quad (diamond shape) or delta (apex up) has the following advantages: it requires only a single support, matches easily into low impedance coax, and offers broadband performance.

In May, 1974, L.V. Mayhead, G3AQC, published a very interesting article on the operation of loop antennas close to ground.[1] By modelling the antennas at UHF, he found that the angle of radiation of a delta loop close to ground could be significantly lowered if it were fed at either side corner instead of at the center of its base leg. This had important and useful implications for the use of such antennas on the lower frequencies.

The corner-fed delta loop has also been mentioned as an effective DX antenna by ON4UN in his book *80-Meter DXing*,[2] and references to it have appeared in many popular Amateur journals over recent years.

Space for a full-sized 7 MHz delta loop was not available at my station in North Sydney, Australia. However, previous experiments carried out in England showed very little deterioration in performance of two-thirds size, side-loaded quads at HF compared to full-sized quads. I decided, therefore, to see whether a corner-fed, reduced-size delta loop for 7 MHz could be made to perform as efficiently as a full-sized loop.

14-MHz model first compared

Instead of experimenting at 7 MHz, I chose to first experiment by reducing the size of a full-sized 14-MHz corner-fed delta loop so that a standard of comparison would be available. The 14-MHz loop had been in use for some time and had shown itself to be an effective antenna despite a base height of only 6 feet (1.8 meters). In comparison tests with a half-wave dipole at 30 feet (9.2 meters), it would generally give a 1 S-point improvement in Europe on long path, and seemed about equal to the dipole on the short path to Europe (from Australia). Both antennas were broadside to Europe on long and short paths.

It is worth remembering that even when the delta loop and dipole delivered equal results, the dipole had the advantage of being nearly a half wavelength high on 14 MHz. To achieve the same effective height for a dipole operating on 7 MHz or 3.5 MHz would mean heights of over 60 feet (18 meters) and 130 feet (40 meters), respectively!

current distribution determines polarization

In the original article on the corner-fed delta loop, the loop was not an equilateral triangle, but instead had sides in the ratio 1:1:1.4, where 1.4 represents the base of the apex-up triangle. This configuration means that the two sloping sides meet at a right angle to each other, and the vertical height of the triangle formed is not as great as it would be if the triangle were equilateral in shape.

The current distribution of a delta loop fed in one corner is shown in **fig. 1**. The phase of the currents in the two sloping legs is such as to make it resemble two vertical antennas fed in phase so that maximum radiation would take place in a plane broadside to the plane of the antennas. Although the sloping sides of the delta loop are at 45 degrees to the horizontal, the

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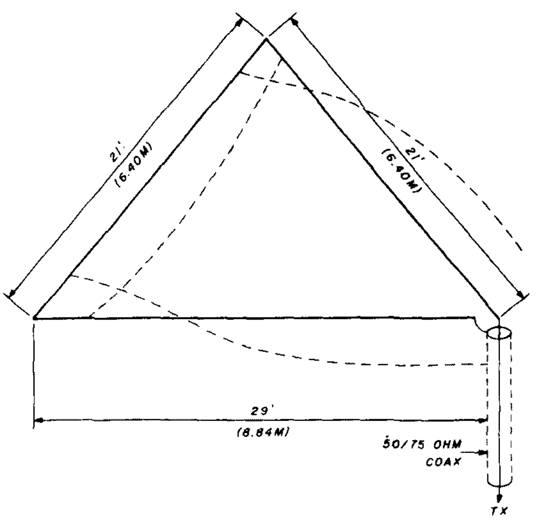

DIMENSIONS FOR FULL-SIZE CORNER-FED DELTA LOOPS

frequency (MHz)	total loop circumference (feet)	(meters)	length of each side (feet)	(meters)	length of base (feet)	(meters)
3.600	279 ft. 2 in.	(85.09)	82 ft.	(24.99)	115 ft. 2 in.	(35.10)
7.075	142 ft. 0 in.	(43.28)	42 ft.	(12.80)	58 ft. 0 in.	(17.68)
14.150	71 ft. 0 in.	(21.64)	21 ft.	(6.40)	29 ft. 0 in.	(8.84)

fig. 1. *Current distribution for a full size 14 MHz corner-fed delta loop. The inner conductor of coax connects to sloping leg.*

phase of the currents in both sloping legs produces a predominately vertically polarized signal.

My objective with the smaller loop was to try to recreate a similar current distribution to that described for the full-sized loop.

A loop two-thirds the size of the normal 14 MHz loop was made. It had sides of 14 feet (4.27 meters), and a base of 20 feet (6.1 meters). However, before I made a serious attempt to load this loop to obtain the conditions of current distribution previously mentioned, I decided to simply series-load the loop with a coil at its feedpoint and observe the effects.

When the loop was loaded in this way, it was possible to lower its resonant frequency from 20.9 MHz to 16.5 MHz by adding small amounts of inductance. However, tuning the loop lower than 16.5 MHz required increasingly larger amounts of inductance, until finally a coil of 17 turns close wound on a 2-inch (5-cm) diameter form was needed for resonance at 14 MHz. This coil had a measured inductance of 17.1 microhenries, indicating that the antenna was 1500 ohms capacitive reactive at 14 MHz. The radiation resistance of the loop was too low to allow a good match into 75-ohm coax, and even with the feeder tapped into the coil to obtain a match, results were very poor. This was not really surprising, because by virtue of being placed at the feedpoint, the coil would have been carrying high current, introducing high loss into the system. The current distribution of the loop loaded in this way would not resemble the full-sized version. At best, then, the loop can be lowered in frequency by up to *20 percent* of its natural resonant frequency by simple series loading with a coil. This is probably not the best way to load the antenna, but it may still give useful results. To bring the 20.9 MHz resonant loop to resonance on 14 MHz represents a *33 percent* lowering of frequency.

the effects of base loading

The next experiment consisted of increasing the sides of the loop from 14 feet (4.27 meters) to 17 feet 9 inches (5.41 meters), but keeping the base at 20 feet (6.1 meters) and to try loading the base wire. If the *base could be loaded, maximum current* would appear in both sloping sides, with a voltage point at the top of the loop. The sides of 17 feet 9 inches (5.41 meters) represent a quarter wave each on the loop at 14 MHz, since the loop circumference in feet is given by 1005/frequency in MHz (bear in mind that there is no *end effect* on the wire of a loop).

The objective with the loop in **fig. 2** was to make the base look like an *electrical half wavelength*, with a voltage point in the middle of the base wire. This would then result in each leg carrying high currents in phase, and a low impedance point at each corner of the loop.

The loading of the base wire was quite easily achieved by connecting two five-foot (1.52-meter) lengths of 300-ohm twin ribbon feeder, shorted at their far ends, 5 feet (1.52 meters) in from each corner of the loop, as shown in **fig. 2**. The ribbon feeder was used instead of a coil *because it was felt losses* would be lower, and ribbon feeder proved easier to trim and *was less bulky than a coil*. A 3-foot 6-inch (1.07 meter) piece of stiff wire was connected at the voltage point midway along the base and brought the loop to resonance at 14.15 MHz.

The loop matched well to 75-ohm coax, with an SWR of less than 1.5:1 across the band. The performance of the loop compared well to the full-size version. However, the two hanging stubs would obviously present a problem when scaled up, so I decided to hang two wires from the middle of the base, and pull them back on themselves, as shown in **fig. 3**. This idea worked extremely well, and with each wire 9 feet 3 inches (2.82 meters) long, the loop was again brought to resonance at 14 MHz, but without the inconvenience of hanging stubs.

This method of loading has the advantages of low loss and ease of trimming. Performance was again similar to the full-sized loop.

reducing leg length

Now that the base had been successfully loaded so that it looked like an electrical half wavelength, an attempt was made to reduce the two sloping legs back to 14 feet (4.27 meters) by taking up the extra wire in the form of a closed stub at the top of the antenna. Various closed stubs using open wire line and 300-ohm ribbon feeder were tried, and although the antenna could be resonated each time, the radiation resistance at the feedpoint was too low to match into 75- or 50-ohm coax.

The stubs were dispensed with, and a single wire 7 feet (2.13 meters) in length connected to the apex of the loop, and hanging vertically brought the antenna to resonance. Although this proved to be a much simpler way of resonating the loop, the radiation resistance at the feedpoint was still too low.

A 4:1 matching transformer at the feedpoint connected in such a way so as to step up the impedance of the loop did provide a match, but the bandwidth of the loop was too narrow. It was felt that the introduction of a matching transformer would start to make the whole exercise rather cumbersome and introduce extra losses apart from the unacceptable bandwidth.

After some experimentation, the sides were increased to 16 feet (4.88 meters), with a 3-foot 6-inch (1.07-meter) wire hanging from the apex. This produced an SWR of less than 1.5 to 1 across the whole band even when the base of the loop was only 4 feet (1.2 meters) high.

Results with this loop were still comparable to the full-sized loop, and represented a reduction of the full-sized loop of 27 percent in terms of circumference, with a corresponding reduction in the perpendicular height of the triangle by 18 percent.

40-meter model

The final experiment was to double the dimensions for 7 MHz operation. This meant that the sides were each 32 feet (9.75 meters), base 40 feet (12.19 meters), apex vertical loading wire 7 feet (2.13 meters), and base loading wires each 18 feet 4 inches (5.59 meters). These dimensions resulted in resonance at around mid-band on 7 MHz with an SWR of no greater than 1.5 to 1 across the band.

A thin, light nylon cord was attached between the bottom of the apex loading wire and the midpoint on the base. This was necessary to keep the apex loading wire from blowing around and also to help prevent sag in the base wire, the latter which was also supporting the bottom loading wires. The loop was pulled slightly away from the mast so as to keep any interaction between the partially metal mast and itself to a minimum.

The results with the 7 MHz loop were good. I worked many continents—including America and Europe — with good reports.

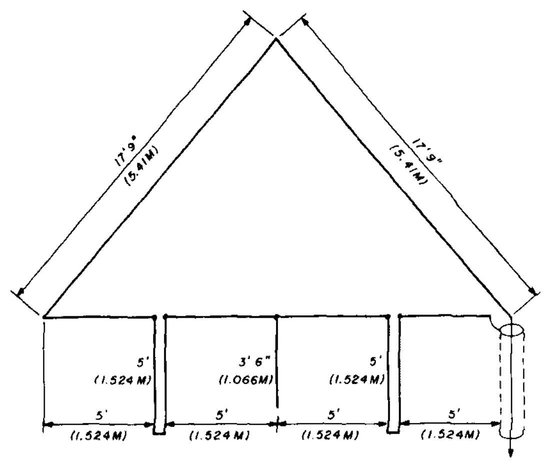

fig. 2. Initial method of loading loop using lengths of 300-ohm ribbon feeder shorted at far end.

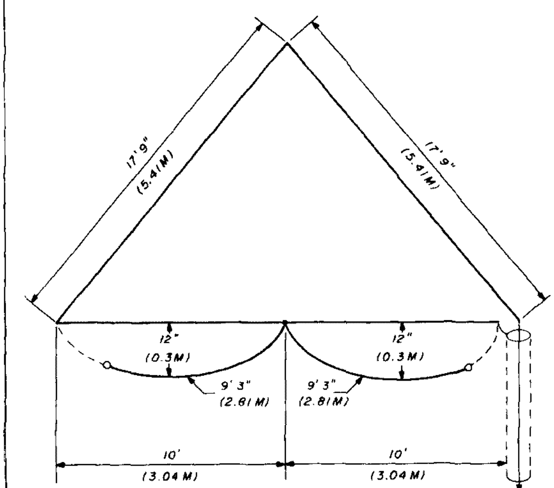

fig. 3. Improved method of loading using wires connected to center of base and pulled out as shown. Light nylon cord is used to secure their ends.

It should be remembered, however, when assessing the performance of any antenna, there is no one antenna that will give excellent results on all paths, during all types of propagation, and over all distances. The *corner-fed delta loop* is a predominately low-angle radiator and as such should generally be at its best over longer distances. If one is only interested in working, for example, up to 1000 miles, a dipole with its higher angle of radiation could be expected to give better results. If a corner-fed delta loop is erected, this point should be remembered in any comparison checks.

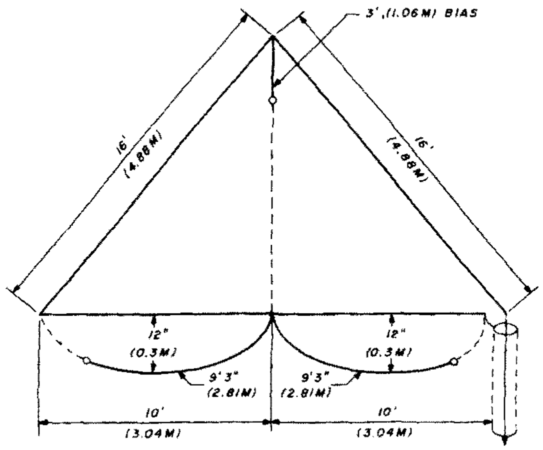

fig. 4. The final loaded loop for 14 MHz. All dimensions are doubled for operation on 7 MHz. Base loading wires are each 18 feet 4 inches (5.59 meters). The top loading wire is made taut by securing to base with thin light nylon cord.

how about 80 meters?

There is no reason those with sufficient space and mast height could not assemble a 3.5 MHz version of this antenna by simply doubling the dimensions given. A 3.5 MHz version would require a mast height of 56 feet (17 meters), which would allow the base of the loop to be 6 feet (1.8 meters) above ground. The horizontal distance taken up would be 80 feet (24.4 meters). These dimensions represent a considerable saving on the full-sized loop and are feasible for many Amateurs.

Final trimming of the loop should be done by adjusting the lengths of the bottom loading wires simultaneously so that their respective lengths are always equal. The top vertical loading wire should not be trimmed. Any trimming should be carried out with the antenna at its normal height, because bringing the base wire closer to ground tends to reduce the resonant frequency of the antenna. The base wire should preferably be a minimum of 6 feet (1.8 meters) above ground. While good results have been achieved with lower positioning than this, it is not recommended; in the original article[1] on the corner-fed delta loop, the base was 10 feet (3 meters) high and this, of course, would be a better height to aim for.

The radiation resistance of a loop antenna drops as it is reduced in size and also as its height above ground decreases. Consequently, 50-ohm coax is recommended for the feeder, although this is not critical. Good results have been obtained with 75-ohm cable.

further experimentation

I hope that by describing some of the experiments carried out, and results obtained, others might be encouraged to experiment further. There is no reason, for example, why those with a little extra space might not erect another similar loop 0.12 to 0.2 of a wavelength *behind* the driven loop, with this additional loop tuned to act as a reflector. To obtain reflector operation of this second loop, the bottom loading wires would have to be increased slightly to bring the loop to resonance some 5 percent lower in frequency from the driven loop.

It should be noted that any attempt to reduce the size of an antenna will be accompanied by a corresponding reduction in radiation resistance, bandwidth, and overall system efficiency. The corner-fed delta loop loaded in this way represents what I believe to be a reasonable compromise between these parameters and acceptable size.

references

1. L.V. Mayhead, G3AOC, "Loop Aerials Close to Ground," *Radio Communication*, May, 1974, page 298.
2. J. Devoldere, ON4UN, *80-meter Dxing*, Communications Technology, Inc., Greenville, New Hampshire, 1978.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

the 5/8-wave VHF antenna revisited

In April, 1935, Gihring and Brown published their classic study on the field strength pattern along the ground for vertical antennas of different heights.[1] Their work was part of an ongoing effort to develop a good anti-fade antenna for the broadcast band. One result of their experiments was the popularization of the 5/8-wave vertical antenna; its design combined high radiation efficiency with a power gain of nearly 3 dB over a 1/4-wave comparison vertical antenna.

Since these classic experiments, the 5/8-wave antenna, in combination with an extensive ground radial system, has become a broadcast industry standard antenna.

Amateurs, however, have used this interesting antenna on the HF and VHF bands with mixed results. When an elaborate ground system is used, the antenna performs as might be expected. But when used with a radial system (as is often the case on the VHF bands), the antenna often proves to be a disappointment.

I found this out last summer when, in an attempt to get into a distant 2-meter repeater, I switched from a 1/4-wave ground plane antenna to a 5/8-wave antenna. Any improvement in signal strength at the far-distant repeater was a product of the imagination.

Why didn't it work? What happened to the 3 dB signal increase I was supposed to get from the bigger antenna? Good question.

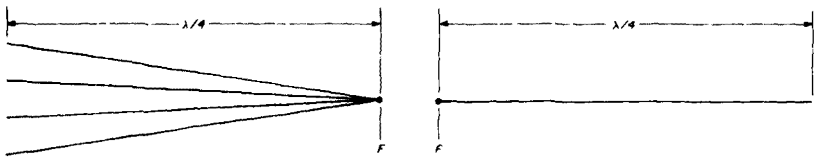

fig. 1. Quarter-wave ground plane, radials collapsed and mounted on its side, transforms itself into a half-wave horizontal dipole.

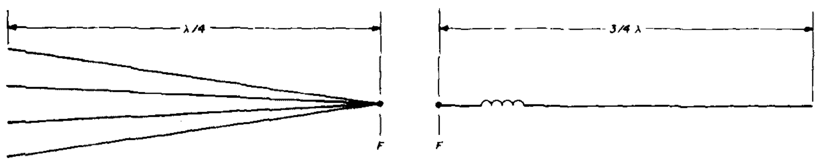

fig. 2. The "5/8-wave" vertical consists of an electrical 3/4-wavelength element and four 1/4-wavelength radials. With radials collapsed and mounted on its side it transforms itself into an asymmetrical antenna.

the W8HXC tests

In my November column I reported on the tests that W8HXC (Ralph) and AF8B (Don) had run on various 2-meter vertical antennas. They found that under many circumstances the feedline became part of the antenna, despite attempts to make the radials isolate the antenna from the feedline, as they are supposed to do.

The simple 1/4-wave ground plane would provide good isolation between antenna and coax line if the line were wrapped into a two-turn RF choke coil about 1 1/2 inches (3.8 cm) in diameter. The coil was placed directly below the antenna. Examining the coaxial line with an "RF sniffer" revealed the line was "cold," provided it dropped directly down beneath the antenna.

The 5/8-wave antenna exhibited current maxima along the outside of the coaxial line until (by cut-and-try) the quarter-wave radials were positioned about 3/8-wavelength below the base of the antenna.

The final conclusion of these tests

was that radials could be placed *any* distance below the antenna as long as the sum of radial length and distance from the antenna base totalled 5/8-wavelength.

the W6SAI tests

This information sounded good to me, so I repeated the tests that Ralph and Don reported. What they said was borne out in fact, but the 5/8-wave vertical antenna still didn't show any appreciable gain over a simple 1/4-wave ground plane antenna.

While I was debating whether I should relegate the antenna to the junk yard, I remembered something I'd read in an old edition of the ARRL *Antenna Book.*[2] The example cited was a horizontal antenna, but the point was made that feeding an antenna in an asymmetrical manner tilted the pattern away from the feedpoint. This could be the clue I was looking for!

If the 1/4-wave ground plane is drawn with the radials in-line with the radiator, it transforms itself into a dipole, as shown in **fig. 1**. If the 5/8-wave vertical is drawn with the usual 1/4-wave radials in-line, it resembles **fig. 2.** The radiator portion of the antenna is really a 3/4-wave resonant section, with 1/8-wavelength wound into a coil so that the advantageous characteristics of the 5/8-wave radiator are retained. The feedpoint (F-F) feeds a lopsided antenna configuration!

Many years ago I had a horizontal "Zepp" antenna that consisted of exactly this configuration (**fig. 3**). *I remember* that it was impossible to balance the antenna currents and RF in the halves of the antenna. I solved the problem by making the short section the same length as the long one. All my loading problems went away.

Enough circumstantial evidence existed at this point for me to substitute 3/4-wavelength radials for the 1/4-wavelength radials on my 2-meter vertical antenna. This is quickly and easily done, as shown in **fig. 4.** This antenna provided improved performance over the 1/4-wavelength ground plane and also over the 5/8-wavelength extended antenna with 1/4-wavelength radials.

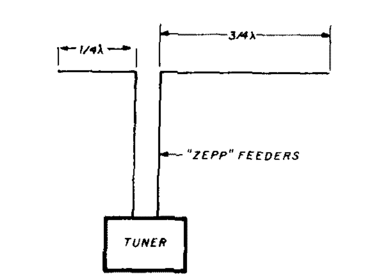

fig. 3. This old "Zepp" antenna at W6SAI looked good on papar but never tuned up properly.

adjusting the 5/8-wavelength vertical

The new antenna was quite easy to adjust. The four radials were precut to 57 inches (145 cm) and the whip was cut two inches longer than estimated length to allow for some pruning. The feedpoint was arbitrarily tapped on the base coil and the antenna was temporarily erected in the air, free and clear of nearby objects.

An SWR check made across the band revealed that the SWR curve canted toward the low frequency end of the band and was quite constant between 1.9-to-1 and 1.7-to-1. The vertical whip was trimmed 1/4-inch (0.6 cm) at a time and the SWR curve checked after each "snip." The SWR seemed to bottom out at about 1.5-to-1, with a very broad response, indicating good bandwidth performance.

Squeezing and expanding the bottom base coil helped a bit and the point of lowest SWR was zeroed in at 146.0 MHz. Unfortunately, the SWR was still too high for us (W6SAI, W6EMD, K6KCM).

The final step was to adjust the feedpoint tap, a quarter-turn along the coil at a time. Now, we were really getting somewhere! After two or three trials, a tap point was found where the SWR on the transmission line was 1.1-to-1, or better, at 146 MHz, rising to about 1.6-to-1 at the band edges.

We observed that the coil tap, number of turns, and vertical antenna length were interrelated. If the antenna was too short, increasing coil induc-

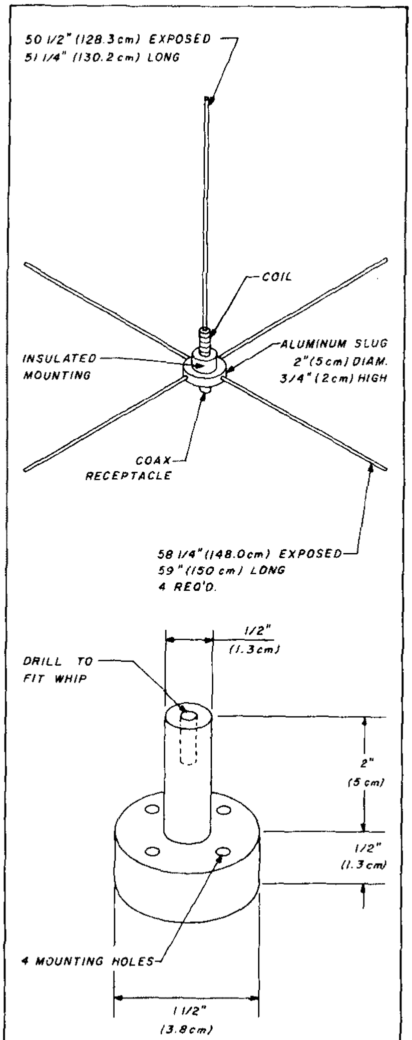

fig. 4. The W6SAI 5/8-wave expanded ground plane for 144 MHz.*

***Note: Elements are built from lengths of 5/16 inch (7.9 mm) diameter aluminum tubing. The vertical whip section is 51-1/4 inches (130.2 cm) long, with 50-1/2 inches exposed above the coil mount. Each radial is 59 inches (150 cm) long, with 58-1/4 inches (148.0 cm) protruding from the aluminum slug. The coil form and support are cut on a lathe from a single block of LEXAN,® a polycarbonate material having good resistance to ultra-violet light (sunlight). It is drilled at the top to accept the whip and at the bottom (4 places) to fit the aluminum mounting slug. The base coil is wound on the top portion of the form. The coil consists of 6-1/2 turns, No. 14 bare copper wire, spaced to 1-1/8 inch (2.86 cm) length. Bottom of the coil is grounded to the aluminum mount. Top end of the coil is attached to the vertical whip by a short strap which encircles the base of the whip and is held in place with 4-40 hardware. Feedline is tapped on the coil 2-1/3 turns from the grounded end. A coaxial receptacle is mounted on the bottom of the *mounting* slug and the whole antenna is mast-mounted by means of an L-shaped bracket.**

band (MHz)	antenna electrical length (half wave)	physical length of radiating element (feet)	resonant frequency (MHz)	2:1 SWR antenna bandwidth (MHz)	ARRL design frequency (MHz)	Amateur band frequency allocation (MHz)
30	5	241	10.105	9.953 to 10.256	10.12	10.100 to 10.15
20	7	241	14.188	13.975 to 14.400	14.17	14.000 to 14.350
16	9	241	18.270	17.995 to 18.544	18.11	18.068 to 18.168
15	11	254	21.210	20.892 to 21.528	21.22	21.000 to 21.450
12	13	254	25.080	24.707 to 25.460	24.94	24.890 to 24.990
10	15	254	28.950	28.523 to 29.305	28.80 29.20	28.000 to 29.700

fig. 5A. Dimensions and specifications of K4EF six-band long-wire array.

tance brought it back into resonance. The whole thing was very forgiving and, when an extra-eager "snip" cut the antenna too short, a slight compression of the base coil brought the SWR back to where it had been before.

No antenna gain measurements were made as no antenna range was available at the time. On-the-air tests indicate that the antenna is doing the job it was intended to do: provide a good signal at distant repeaters that could not be triggered with a conventional ground plane antenna.

who was first?

Amateur Radio historians often come up with unusual and interesting items. The 5/8-wave vertical antenna popularized by Gehring and Brown in 1935, was patented in the United States by Andrew Alford on April 2, 1940. However, the same antenna concept had been patented in Germany on October 24, 1932, and even this was an extension of a previous patent on the same antenna design issued earlier in Russia.

To top it off, the popular coaxial balun (so-called "Collins Balun") was patented by W.B. Bruene with U.S. patent 2,777,996 in 1954. An identical balun was patented in Germany by Telefunken in August, 1942.

It looks as if a great deal of effort has been expended in re-inventing the wheel!

My thanks to Dipl. Eng. A. Krischke, DJ0TR, OE8AK, for these details. This well-known historian has a private collection of over 1000 radio patents dating back to the early antenna patents of Marconi (1896), Lodge (1898), and Braun (1898).

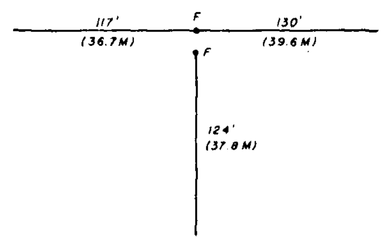

fig. 5B. Top view of K4EF antenna array.

the K4EF antenna for 10-30 MHz

This simple antenna has consistently outperformed a straight long-wire antenna on long distance DX contacts. It was developed by Ev (K4EF) to cover the new ham bands at 18 and 24 MHz, as well as the regular 30, 20, 15, and 10 meter bands. The feedpoint impedance is close to 200 ohms on all bands, so a good quality 4-to-1 balun provides a convenient match to a 50-ohm transmission line. No antenna tuner is necessary, except on the low (CW) end of the 10-meter band.

Antenna dimensions are given in **fig. 5A**. A view of the array is shown in **fig. 5B**. Wire length is critical and should be duplicated to within 2 inches (5 cm). Copper-clad steel wire is recommended, as pure copper wire will stretch over such a span. If kept reasonably in the clear, away from large metal objects, and thirty feet (9 meters) or higher above ground, the electrical characteristics will closely match those listed in the illustration. The antenna may be mounted on a mast or tower at the feedpoint and the three legs suspended in a flat-top or inverted-V configuration. The apex angles of the wire legs are not critical and the sum of the angles may be varied from 180 degrees to 120 degrees, or less.

lightning protection

Lightning is a problem in many areas of the country where frequent electrical storms occur. This antenna is no more vulnerable than other antennas of its size and it has the advantage of being easily protected. The balun (which can be destroyed by a nearby lightning strike) and coax are moved to the base of the tower and the antenna is fed with a 200 ohm open-wire line which can be switched to ground at the tower base when the antenna is not in use (**fig. 6**.)

The line is made up of parallel-connected lines. K4EF's line used two insulators: one at the top and the other at the bottom. Thirty-pound weights were attached to the bottoms of each wire to keep the whole line under tension. This eliminated intermediate insulators. Surplus, heavy-duty 208 ohm "ribbon" line can also be used for the transmission line.

receiver overload

The large capture area of the antenna results in large signal voltages (sometimes from unwanted stations) which may overload the front end of

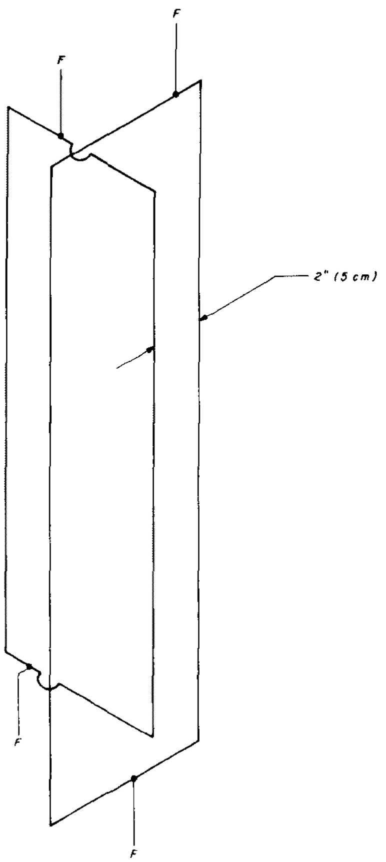

fig. 6. Construction of a 200-ohm line. Diagonally opposite wires are joined at each end of line. Spacing to centers of wires on both sides of square is 2 inches (5 cm) for No. 10 wire. Base of line can be grounded when antenna is not in use.

the receiver. A simple RF attenuator will help, if your receiver does not have some sort of input protection. The circuit of **fig. 7** is suggested.

protecting wooden masts at or below ground level

Fourteen years ago my friend Stu, W2LX, erected a wooden mast on a ground post sunk into moist soil. The post was a 4 × 4, about five feet long, with over 3 feet sunk into the ground. The problem was how to protect the portion of the post buried in the ground.

A previous mast had been treated heavily with wood preservative, but it has rotted out after a few years in the ground. This time, Stu had a better idea.

An untreated mast post was wrapped with three layers of aluminum foil — the heavy-duty type used in the kitchen. The layers of foil were arranged so that the seams did not overlap, and the foil was carefully folded around the base of the ground post. The foil was wrapped with vinyl tape at several points. The mast was then placed in the ground hole and the hole filled with dirt and tamped down.

The metal foil protruded a few inches above ground to protect the post from surface water.

This fall — *fourteen years later* — W2LX moved to a new location. Because the new owner of his home would have no use for the mast, it was taken down. When the ground post uprooted, Stu found the unusual protective technique had been an unqualified success! Once the soil had been washed from the post, the foil looked practically new. Carefully removing the tape and unwrapping the foil, he found that the wood *also* looked almost new. There was no sign of water damage, termites, or rot.

W2LX plans to use this technique when he erects his wooden mast at his new QTH and passes the idea along to other Amateurs who may be interested in erecting a wooden mast.

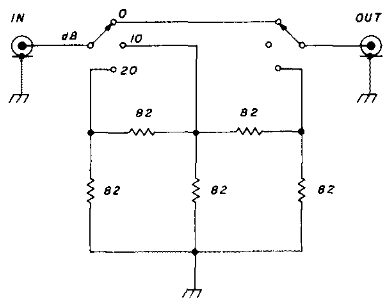

fig. 7. Simple two-step attenuator.

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ham radio

high power RF switching with pin diodes

Recent developments in PIN diode technology allow fast, silent switching at 1500 watts+

The PIN diode is a semiconductor device that operates in a manner similar to that of a variable resistor at RF frequencies. The amount of forward DC bias applied to the PIN diode determines its resistance (impedance) to the passage of RF signals.

PIN diodes are attractive for RF switching because they have no moving parts; can "hot switch" large RF currents; can control large amounts of RF current with a relatively small DC bias control current; and don't introduce any significant RF waveform distortion.

PIN diodes are formed from three distinct types of silicon wafers: an *I*ntrinsic layer (pure, non-doped); a *P* doped layer, and an *N* doped layer. **Figure 1** shows a typical PIN diode with its layers identified. A practical package for a PIN diode has the leads attached to the P and N layers and the whole unit encapsulated in either epoxy or glass. It is the thickness of the intrinsic layer which determines the "geometry" of the PIN diode and gives the diode manufacturer the ability to create PIN diodes with different characteristics for special applications.

The external physical appearance of PIN diodes is determined by their intended applications. **Figure 2** shows two Unitrode PIN diodes: a 7300 series and a 4000D series. The 7300 series, about the same size as a 1N4148, is used for microwave attenuators because of its low internal capacitance (0.7 pF). The larger one, series 4000D, is an insulated stud mounted unit with ribbon leads. This unit is used for high power RF switching and certain applications at a 500 *Kilowatts* pulsed power level (1 μs pulse). Tests have shown that with proper heatsinking and DC biasing, the Unitrode 4000D series PIN diode can handle in excess of 3000 watts.*

PIN diode parameters and specifications

All PIN diodes, regardless of their application, share certain common characteristics. One is a forward resistance, R_s, that varies inversely with DC forward bias. This is usually shown in graphical form. (See **fig. 3**, a graph for a typical Unitrode UM4000 series PIN diode.) Note that at 1000 mA (1 amp) forward bias, the UM4000 R_s is approximately 0.1 ohm. At 1 mA the R_s rises to 20 ohms and at 1 μA forward bias, the R_s is in excess of 10 kilohms. The standard for rating most PIN diodes is to provide 100 mA forward bias current at 100 MHz.

Figure 4 shows a set of equivalent circuits for PIN diodes in both a forward and reverse bias state. The forward-biased PIN diode can be considered as equivalent to a series resistor (R_s) and inductor (lead inductance). The reverse-biased PIN diode is equivalent to a resistor (R_p-parallel resistance) in parallel with a capacitor (the C_T . . . total package capacitance) with both of these in series with an inductor (the

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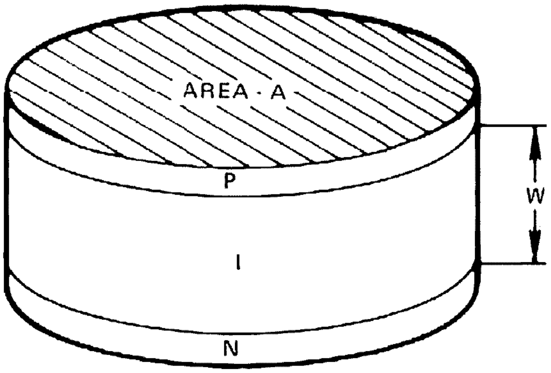

fig. 1. PIN diodes consist of three distinct layers: *P* type, *N* type, and a pure non-doped (*I*ntrinsic) layer.

leads). When PIN diodes are reverse biased, they exhibit a blocking effect to RF as shown in the equivalent circuit of **fig. 4**. As the reverse DC bias is increased, R_p increases and C_T decreases. The limit of the reverse DC bias is determined by the value of V_r (reverse breakdown voltage) of the particular diode. **Figure 5** shows a typical curve of reverse bias voltage (V_r) versus parallel resistance (R_p) for a Unitrode 7300 series PIN diode. **Figure 6** shows a typical curve of reverse bias voltage (V_r) versus total capacitance (C_T) for a Unitrode 4300 series PIN diode.

power handling capability

The maximum power rating of a PIN diode is a function of the forward resistance, R_S, and the amount of RF current flowing through the diode. A PIN diode rated to dissipate 12 watts at 25 degrees C can safely switch 1500 watts of RF. A typical calculation shows why this is so. Assume:

$$\begin{aligned} &RF\ load &&= 50\ ohms \\ &R_s &&= 0.2\ ohm \\ &RF\ power\ level &&= 1500\ watts \end{aligned}$$

$$I = \sqrt{P/R} = \sqrt{\frac{1500}{50}} = 5.48\ amperes$$

The power dissipated by the diode is equal to forward resistance times current squared or:

$$0.2 \cdot (5.48)^2 = 6\ watts$$

The power dissipation of a PIN diode is therefore a function of the load impedance, the forward resistance and the RF current flowing. Consequently, a PIN diode rated to dissipate only 12 watts can, with proper heat sinking, easily handle RF in excess of 1500 watts.

PIN diodes versus vacuum relays

The use of vacuum relays for "high speed" RF

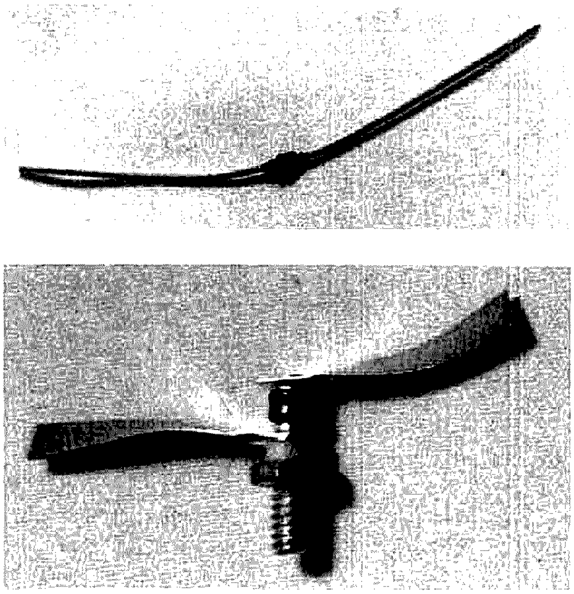

fig. 2. One of the PIN diodes is used in microwave attenuators while the larger unit can actually handle 500 kW of pulsed power.

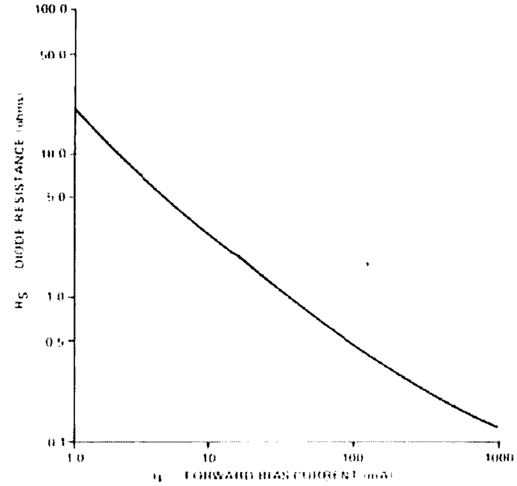

fig. 3. R_s versus current for UM4000 PIN diode.

switching has been well covered in the literature over the past 20 years (see bibliography). Several commercial amplifiers use vacuum relays for RF switching in order to obtain full break-in "QSK" operation. But using vacuum relays to obtain the high-speed switching times required for full QSK has several disadvantages. These include high cost; the inability to "hot switch"; the necessity for complex switching and protective circuitry; mechanical sound and vibration as the relay opens and closes on each character of CW

[1] Tests performed at Design Electronics Ohio, Groveport, Ohio.

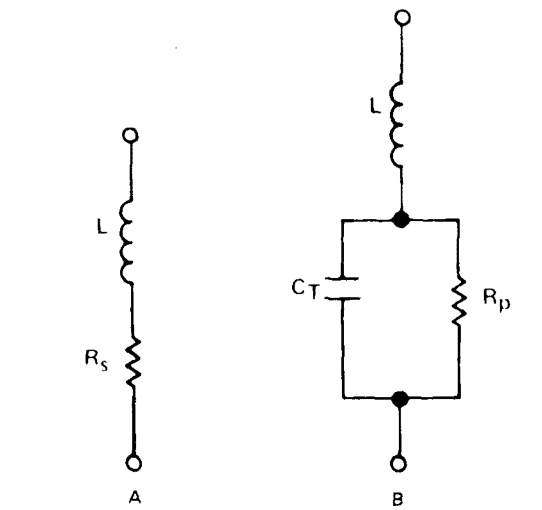

fig. 4. PIN diode equivalent circuits. (A) forward bias, and (B) reverse bias.

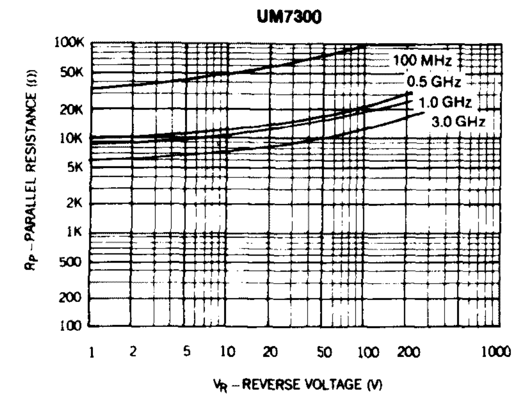

fig. 5. Parallel resistance of a Unitrode UM7300 PIN diode versus reverse voltage over the 100 MHz to 3 GHz frequency range.

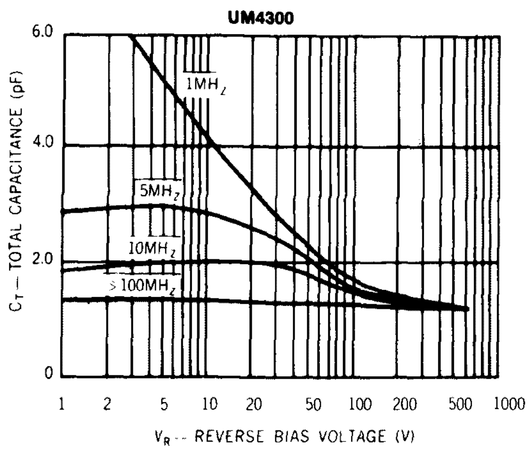

fig. 6. Unitrode 4300 series PIN diode capacitance versus reverse voltage response.

or AMTOR; and maximum switching time of approximately 1-2 milliseconds. Amateur use of PIN diodes for RF switching is a relatively recent occurrence. However, nearly all currently available transceivers that offer full QSK use PIN diodes for their T/R function. Several articles have appeared in the literature over the past 10 years concerning the use of PIN diodes in "low power" T/R switching. Using PIN diodes to perform RF switching functions offers several advantages, including relatively low cost; the ability to be "hot switched"; silent operation; rugged construction; and no complex protective or peripheral circuitry are required. In addition, switching times less than 1 μs are possible.

making PIN diodes work at HF

PIN diodes were designed for and operate best in the VHF and UHF regions. Their use below 30 MHz was delayed because of the need for high inductance and high current RF chokes. The need for capacitors capable of handling 5-10 amps of RF current, with values up to 100,000 pF was also a barrier. PIN diodes can be used at HF, however, and a detailed description of a commercial application that successfully utilizes PIN diodes for RF switching follows.

the QSK 1500

The QSK 1500 switch was developed to allow owners of QSK transceivers to operate full break-in CW or AMTOR at the legal power limit (1500 watts). It uses pin diodes to provide ALL the T/R switching functions associated with the relays in the existing amplifier, with no modifications to either the QSK transceiver or the RF amplifier needed. However, this unit is not intended to make a non-QSK transceiver operate in the QSK mode, nor will it work with a separate transmitter/receiver combination.

A block diagram of the QSK 1500 is shown in **fig. 7** and the schematic of the RF switching section in **fig. 8**.

receive signal path

The receive signal from the antenna travels through the OUTPUT RECEIVE LINE BLOCKER, the RECEIVE LINE PROTECTOR, and the INPUT RECEIVE LINE BLOCKERS, then into the front end of the QSK TRANSCEIVER. With the QSK

1500, the "receive signal" never passes through the RF amplifier, but is instead always bypassed around it by the PIN diodes in the QSK 1500. The receive signal from the antenna, is prevented from seeing the tank circuit of the RF amplifier by PIN diode CR2, which is reverse biased. This diode prevents "suck out" or attenuation of the receive signal. PIN diodes CR3, CR4, and CR5 are forward biased and offer a very low impedance path for the receive signal to reach the QSK transceiver. Capacitors C1 and C4 are selected so that receive signal attenuation is typically less than 0.5 dB.

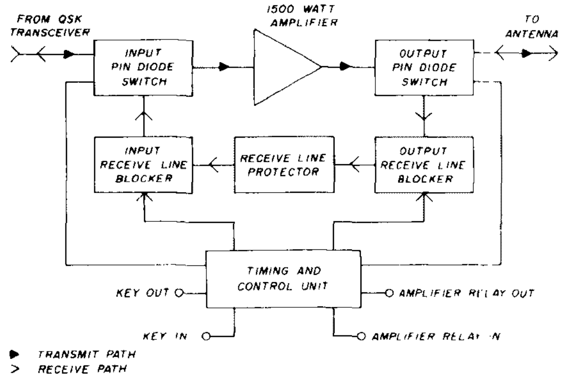

fig. 7. Block diagram of QSK 1500.

transmit signal path

The sequence of events is more complicated in the transmit mode. Let's follow the progress of two transmitted "dots" of CW. (The same pattern also occurs for AMTOR.)

The initial key closure causes the following to happen:

- The timing circuit is triggered on the control board. (This timer is a retriggerable one-shot with a time out of 1.32 seconds.)
- The AMP RELAY OUT line closes and the relays in the RF amplifier are activated.
- The INPUT AND OUTPUT RECEIVE LINE BLOCKERS (PIN diodes CR3, CR4, CR5) are reverse biased with 525 volts DC.
- The input and output PIN diodes switch (CR1 and CR2) are forward biased. CR1 (the input PIN diode) is forward biased with 125 mA DC current, while CR2 (the output PIN diode) is forward biased with 950 mA DC current.
- The KEY OUT LINE from the QSK 1500 triggers the CW KEY JACK on the QSK transceiver and a "dot" of CW is generated.
- The RF from the QSK transceiver flows through C1,

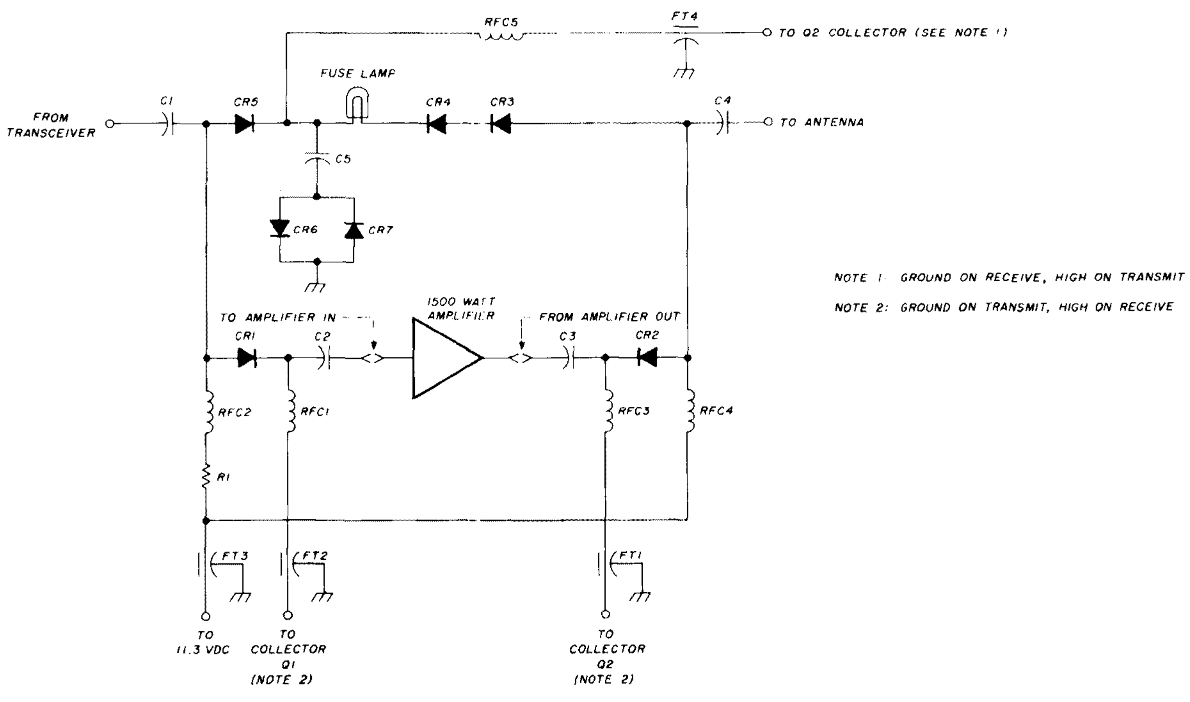

fig. 8. Schematic of RF switching section of QSK 1500.

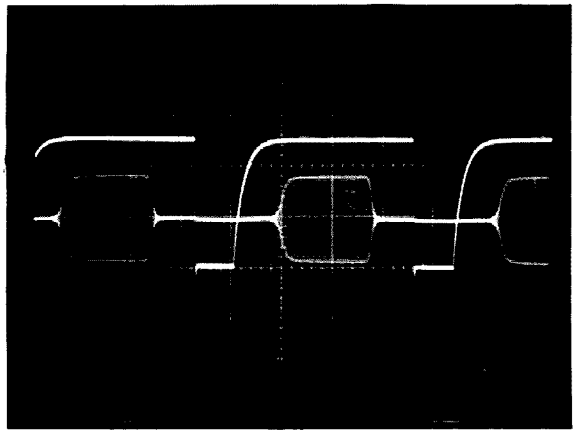

fig. 9. CW waveform inside 500V blocking voltage. (See Note below.)

CR1, and CR2 into the RF amplifier, where it is amplified and then passed through C3, CR2, and C4 into the antenna.

- The 525 volt DC reverse bias applied to PIN diodes CR3, CR4, and CR5 prevents any transmitted RF signal from passing through the RECEIVE LINE.

As soon as the "key" opens the following occurs:

- PIN diodes CR1 and CR2 switch state and go from a forward bias condition to a reverse bias condition.
- PIN diodes CR3, CR4, and CR5 also switch state and go from a reverse bias state to a forward state.
- With the reversal of the bias states of all five PIN diodes, we have now returned to the receive condition.

Figure 9, an oscilloscope photo of a CW waveform (58 WPM dots) inside the 500 volt blocking voltage, shows the "turn on"/"turn off" of the PIN diodes in relation to the blocking voltage.

With every key opening and closure a change of state of the PIN diodes occurs. The relays in the RF amplifier *remain* closed until no key closure has occurred for 1.32 seconds. This assures that no RF is being switched by the RF amplifier's internal relays, and that all switching between transmit and receive is accomplished by the PIN diodes in the QSK 1500.

The RECEIVE LINE PROTECTOR ensures that no RF can reach the transceiver or the RF amplifier input in the unlikely event of an output blocker PIN diode (CR3 or CR4) failure.

Note: Photo shows transmission of a string of dots at 58 WPM. At this rate the receive time between pulses is 8 ms. The built-in delay time between turn-on of the 500V blocking voltage and the beginning of the CW envelope is 7 ms. The CW pulse (dot) is 21 ms in duration; the built-in delay between the end of the CW pulse and turn off of the 500V blocking voltage is 7.5 ms. The CW waveform shows no trace of distortion. Power level, 1380 watts; transmitter, TS930S; amplifier, Drake L7; QSK unit, QSK 1500; scope, TEK 556 dual-trace, wattmeter, Bird 43 with scope coupler; load, Bird Termaline™ 2 kW; keyer, Accu-keyer II. Vertical scale = 200V/cm (× 10 probe with 20V scale); horizontal scale = 10 ms/cm.

conclusion

The appearance of high power PIN diodes represents the dawn of a new era in RF switching. The vacuum relay — the former "king of RF switching" — will slowly but surely give way to the solid state PIN diode. PIN diodes now make possible relatively low-cost, ultra high-speed RF switches, which if properly designed and constructed can operate at power levels exceeding 1500 watts. A totally silent switch in a small package, the PIN diode is here to stay.

acknowledgement

Thanks are due to the technical and engineering staff at Unitrode for assistance with technical data and help with solving various problems that arose.

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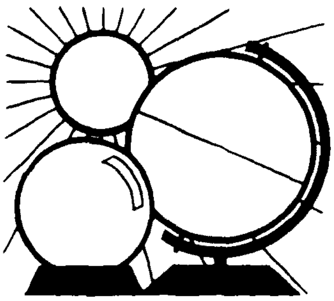
ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

ionosphere matching

Over the years a few hams have tried to vary the height of horizontal antennas above ground in order to vary the take-off-angle (TOA). By varying the angle at which the signal reflects from the ionosphere, you can control the length of the hops (and consequently, distance) and by orienting the antenna (rotating the beam), the direction the signal takes. This is known as "matching to the ionosphere."

If you can't vary your antenna's height above ground, then fix it permanently at a height that emphasizes your area of interest. (Antenna handbooks include graphs that relate take-off angle, ionospheric layer height, MUF and path length.)* **Table 1** shows a short BASIC program for calculating the take-off angle of a horizontal antenna given the height above ground.

table 1. BASIC program for calculating take-off angle of a horizontal antenna given height above ground. (This program was written for the IBM 4341 in IBM's BASIC/VS.

```
110 R1=180/&PI
120 REM GET INPUT DATA
130 PRINT "ANTENNA HEIGHT ABOVE GROUND, FEET"
140 INPUT H1
150 PRINT "FREQUENCY TO BE USED, MHZ"
160 INPUT F1
170 REM CALCULATE A FULL WAVELENGTH, L1, AT THIS FREQ.
180 L1=984/F1
190 REM CALCULATE FIRST LOBE ANGLE, A1
200 A1=ASN(L1/(4*H1))
210 A1=A1*R1
220 PRINT "ANGLES OF MAXIMUM RADIATION FROM THIS ANTENNA"
230 PRINT
240 PRINT "FIRST LOBE ANGLE, DEG.=",A1
250 PRINT
260 REM CALCULATE 2ND LOBE ANGLE, A2
270 IF (4*H1)<(3*L1) GO TO 500
280 A2=ASN((3*L1)/(4*H1))
290 A2=A2*R1
300 PRINT "SECOND LOBE ANGLE, DEG.=",A2
310 PRINT
320 REM CALCULATE 3RD LOBE ANGLE, A3
330 IF (4*H1)<(5*L1) GO TO 500
340 A3=ASN((5*L1)/(4*H1))
350 A3=A3*R1
360 PRINT "THIRD LOBE ANGLE, DEG.=",A3
370 PRINT
380 REM CALCULATE 4TH LOBE ANGLE, A4
390 IF (4*H1)<(7*L1) GO TO 500
400 A4=ASN((7*L1)/(4*H1))
410 A4=A4*R1
420 PRINT "FOURTH LOBE ANGLE, DEG.=",A4
430 PRINT
500 PRINT "NO OTHER SIGNIFICANT LOBES"
510 END
```

last-minute forecast

The high probability of increased solar flux in the middle of January makes the second and third weeks of the month *favorable for* DX openings on the 10 through 30-meter bands. The openings may be transequatorial in nature, particularly if minor geomagnetic *field disturbances occur at* the same time as this greater solar activity. The lower frequency bands should be good during the first and last weeks of the month. Low noise and signal absorption during low solar flux periods account for good daytime openings on 40 and 80 meters this year. Geomagnetic field disturbances that produce ionospheric high latitude trough conditions (see December, 1984, *DX Forecaster*) should not be too significant this year except for periods around January 2, 11, 21 and 29.

Lunar perigee occurs on the 15th, with a full moon on the 7th this month. An intense but short-duration meteor shower, the Quadrantid, will reocccur between January 2nd and 4th and last a few hours.

band-by-band summary

Ten and fifteen meters will be open worldwide from sunrise until after sunset during the solar flux peaks this month. Skip distances of 2500 miles (4000 km) (or multiples) are possible, and will occur on the daylight paths.

Twenty meters will be open to some area of the world for the entire 24-hour period many days of the month with the band conditions peaking in all directions just after local sunrise and again toward the east and south during the late evening hours. During hours of darkness the band will peak toward the west in an arc from southwest through northwest, encompassing the Pacific areas.

Thirty meters is a daytime and nighttime band. During the day it will re-

*Next month's *DX Forecaster* will feature such a graph.

JANUARY

WESTERN USA

GMT	PST	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)
0000	4:00	30	30	15	15*	15	10	10	20
0100	5:00	20	30	15	10	15	10	10	20
0200	6:00	20	30	15	10	15	10	10	20
0300	7:00	20	30	20	10	15	15	10	20
0400	8:00	20	30	20	10	15	15	15	20
0500	9:00	20	30	20	10	15	15	15	20
0600	10:00	20	40	20	15	20	20	15	30
0700	11:00	30	40	20	15	20	20	20	30
0800	12:00	30	40	20	15	20	20	20	30
0900	1:00	30	40	20	20	20	20	20	30
1000	2:00	30	40	20	20	20	20	20	30
1100	3:00	30	40	30*	20	20	20	20	30
1200	4:00	30	40	20	20	20	20	20	30
1300	5:00	30	40	20	20	20	20	20	40
1400	6:00	40	30	20	20	20	20	20	40
1500	7:00	40	30	15	20	20	20	20	40
1600	8:00	40	30	15	20	20	20*	20	40
1700	9:00	40	20	15	20	15	15	20	40
1800	10:00	40	20	15	15	15	15	15	30
1900	11:00	40	20	15	15	15	15	15	30
2000	12:00	40	20	15	15	15	15	15	20
2100	1:00	30	20	15	15	15	15	15	20
2200	2:00	30	20	15	15	15	15	15	20
2300	3:00	30	30*	15	15	15	15	15	20

MID USA

GMT	MST	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)	CST
0000	5:00	30	30	15	15	15	15	10	20	6:00
0100	6:00	30	30	15	20	15	15	10	20	7:00
0200	7:00	40	30	20	20	15	15	15	20	8:00
0300	8:00	40	30	20	20	15	15	15	20	9:00
0400	9:00	40	30	20	20	20	15	15	20	10:00
0500	10:00	40	30	20	20	20	20	20	30	11:00
0600	11:00	40	40	20	20	20	20	20	30	12:00
0700	12:00	40	40	20	20	20	20	20	30	1:00
0800	1:00	40	40	20	20	20	20	20	30	2:00
0900	2:00	30	40	20	20	20	20	20	30	3:00
1000	3:00	30	40	20	20	20	20	20	30	4:00
1100	4:00	30	40	20	20	20	20	20	30	5:00
1200	5:00	30	40	20	15	20	20	20	30	6:00
1300	6:00	30	40	20	15	20	20	20	40	7:00
1400	7:00	20	30	15	15	20	20	20	40	8:00
1500	8:00	20	30	15	15	20	20	20	40	9:00
1600	9:00	20	30	15	15	20	15	20	40	10:00
1700	10:00	20	20	15	10	15	15	20	40	11:00
1800	11:00	20	20	15	10	15	15	15	30	12:00
1900	12:00	20	20	10	10	15	15	15	30	1:00
2000	1:00	30	20	10	10	15	15	15	20	2:00
2100	2:00	30	20	10	10	15	15	15	20	3:00
2200	3:00	30	20	15	10	15	15	15	20	4:00
2300	4:00	30	30	15	15	15	15	15	20	5:00

EASTERN USA

GMT	EST	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean, S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)
0000	7:00	30	30	15	15	15	15	15*	20
0100	8:00	30	30	15	20	15	15	15	20
0200	9:00	40	30	20	20	15	15	15	20
0300	10:00	40	30	20	20	15	15	15	20
0400	11:00	40	30	20	20	20*	20	15	30
0500	12:00	40	30	20	20	20	20	20	30
0600	1:00	40	40	20	20	20	20	20	30
0700	2:00	40	40	20	20	20	20	20	30
0800	3:00	40	40	20	20	20	20	20	30
0900	4:00	30	40	20	20	20	20	20	30
1000	5:00	30	40	20	20	20	20	20	30
1100	6:00	30	40	20	20	20	20	20	40
1200	7:00	30	40	20	15	20	30*	20	40
1300	8:00	20	20	15	15	20	20	30*	40
1400	9:00	20	20	15	15	20	20	20	40
1500	10:00	20	20	15	15	20	20	20	40
1600	11:00	20	20	15	15	20	15	20	40
1700	12:00	20	20	15	10	15	15	15	40
1800	1:00	20	20	15	10	15	15	15	30
1900	2:00	30	20	15	10	15	15	15	30
2000	3:00	30	20	10	10	15	15	15	20
2100	4:00	30	20	10	10	15	15	15	20
2200	5:00	30	20	15	15	15	15	15	20
2300	6:00	30	20	15	15	15	15	15	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

semble 20 meters, although signal strengths may decrease during midday during days of high solar flux values. This band will also be useful well into the night and often throughout the night. Once again exceptions to this are nights that follow very high solar flux value days. The poor period is usually the hour or so before dawn (diurnal MUF minimum). The workable distance may be expected to be greater than that of 80-meter DX at night and less than that of 20 meters during the day.

Forty and eighty meters will be the most useful nighttime DX bands. Forty during the daylight hours will be like 30 meters, with lower midday signal levels, *no* predawn propagation failure, and shorter skip distances overall. At night most areas of the world will be workable from slightly before dusk until a little after sunrise. Hops shorten on these bands to about 2000 miles (3218 km) for 40 meters and 1500 miles (2414 km) for 80 meters, but the number of hops can increase because signal absorption in the D-region of the ionosphere is low during the night. The path follows the direction of darkness across the earth, similar to the way in which the higher bands follow the sun.

One-sixty meters will be similar to 80 meters, providing good working conditions for enthusiastic DXers who like to operate during the night and early morning hours, especially at local dawn.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

VHF/UHF high power amplifiers: part 1

Until August 28, 1983, FCC Amateur Radio regulations limited transmitters to a maximum of 1000 watts DC plate input power on CW or 2000 watts PEP input on SSB. This was in contrast to most commercial radio regulations, which are specified in terms of output power. The regulations also specified that any drive power passed on to the amplifier output (such as in grounded grid or transistor amplifiers) had to be accounted for by lowering the DC input power to the final amplifier stage accordingly.

These archaic regulations particularly penalized VHF/UHF experimenters who wanted to take advantage of the maximum power limit for DX and for specialized communication modes such as EME. Klystron and transistor UHF amplifiers frequently have only 35 percent efficiencies; most UHF high power tube amplifiers have typical efficiencies of only 50 to 60 percent. Therefore, when the legal power limits were required, higher priced vacuum tubes were usually necessary in order to maximize efficiency — *often at the expense of linearity*!

These regulations are now history. On August 29, 1983 the FCC revised Part 97 of the Amateur Radio Rules and Regulations, authorizing the measurement of power in terms of maximum PEP (peak-envelope-peak) output, with 1500 watts the maximum limit (except in special subbands of the spectrum where other concerns may necessitate a lower power limit). SSB operation is not particularly affected because a 55 percent efficient VHF amplifier under the old regulations would have delivered up to 1100 watts of PEP output, only 1.35 dB down from 1500 watts. However, a 60 percent efficient CW amplifier would be, under the old regulations, about 4 dB below the new 1500 watts PEP output (CW and PEP output are the same power)!

If you want to run the new power limit in an Amateur VHF/UHF power amplifier, these regulatory changes have profound implications for the design of your next amplifier. Because most articles in the Amateur literature pertaining to VHF/UHF power amplifier design were written before the regulations were changed, this month's — and next month's — column address the broad subject of Amateur VHF/UHF high power amplifiers, with special emphasis on the effects of the new FCC regulations. (Many of the topics discussed will also apply to HF.) A list of recommended references will also be included so you can be sure your amplifier will be equal to the state-of-the-art. I'll concentrate on vacuum tube amplifiers because the typical solid-state devices available today can't economically furnish the maximum legal power in the VHF/UHF spectrum.

power tubes in general

Most high-power transmitting tubes used by Amateurs are technically classified as power grid tubes. Although some pentodes are used on 6 meters, most VHF/UHFers primarily use triode and tetrode power grid tubes. Some of the most popular types are listed, with some of their important specifications, in **table 1**.[1]

Triodes are most often used when a single high voltage supply is desired. With triodes, circuitry is usually less complicated than with tetrodes, especially if a cathode-driven grounded-grid circuit is used. Gain is low to moderate in this type of configuration, but the amplifier is usually easier to stabilize. Tetrodes, on the other hand, require a more complex high voltage supply with a screen grid bias voltage and may require neutralization. However, they usually have higher power gain than triodes, especially if grid-driven.

Most modern power tubes come in either a glass envelope with an internal anode or a ceramic package with an external anode. (The latter are more prevalent on 2 meters and above.) The

table 1. Popular VHF/UHF power amplifier tubes and typical parameters listed in order of plate dissipation.

types	dissipation[1]	maximum power[2]	PEP[3]	IMD[4]	F_{MAX}[5]	F_{EXT}[6]	other notes
triodes							
8875	300	770	590	−35	500	900	cathode driven
3CX400A7/8874	400	770	590	−35	500	900	cathode driven
3-500Z	500	1400	740	−40	110	-	cathode driven
3CX800A7	800	1320	750	−36	30	450	cathode driven
3-1000Z/8164	1000	4800	1080	−29	110	-	cathode driven
3CX1500A7/8877	1500	4000	2050	−38	220	400	cathode driven
8938	1500	4000	2030	−44	500	-	cathode driven
tetrodes							
4X150A (old)	150	375	200	-	500	-	discontinued
4X150A/7034	250	500	200	-	150	500	new style anode
4CX250B/7203	250	500	295	−25	500	-	
4CX250R/7580W	250	500	295	−25	500	-	"Ruggedized"4CX250B
4CX300A/8167	300	500	350	−25	500	-	
4CX350A/8321	350	750	300	−30	30	220	poor efficiency at UHF
8930	350	600	350	−27	500	-	formerly DX393
8122	400	660	380	−29	500	-	
4-400A/8438	400	1400	495	−35	110	-	
7650/7651[7]	600	1500	680	−31	1215	-	
4-1000A/8166	1000	4200	1540		110	-	
4CX1000A/8168[8]	1000	3000	1400	−23	110	220	
4CX1000K	1000	3000	1400	−23	110	400	Improved in/out isolation
7213/7214[7]	1500	2500	1250		1215	-	
GL6942	1500	2800			900	-	cathode driven
4CX1500B/8660	1500	2700	1160	−43	110	220	

Notes:
1. Rated plate power dissipation in watts if adequately cooled.
2. Maximum CCS DC input power.
3. Maximum useful peak envelope output power in watts at a low frequency, typical 2-30 MHz.[2]
4. Typical third-order intermodulation distortion level at rated PEP indicated, calculated in 2-30 MHz region.[2]
5. Upper frequency at which the maximum ratings apply.
6. Upper useful frequency if plate voltage and input power are reduced.
7. The second tube listed is a pulsed rated version. It will probably perform the same as the first version.
8. No grid current allowed.

older glass-sealed transmitting tubes with external anodes (for example, the 4X150A) were very efficient *if operated within their ratings* but were not very rugged. However, the newer ceramic insulated tubes, though perhaps somewhat less efficient, will usually take more severe punishment.

Transmitting tubes are usually designed for a particular mode of operation such as Class C, linear, or pulse. Class C operation is preferred for CW operation when high efficiency is required. Linear operation is required for SSB operation. Amateur Radio pulse, a form of Class C, is not permitted below 2.3 GHz.

The tubes designed for Class C operation typically have high efficiency in Class C but will not always be very linear when biased for linear operation, especially when operated near their maximum ratings. Furthermore, Class C amplifiers usually have lower power gain than equivalent linear amplifiers and often generate key clicks.

A good example of a frequently misused VHF/UHF Class C tube is the popular 4CX250B tetrode. This tube was very popular under the old FCC regulations of 1000-watts input power because a pair could be operated efficiently (at 60 to 70 percent efficiency) through 70 cm, a requirement for most EMEers. However, they are also often used in linear service. At the maximum rating of 500 watts input power per tube, the IMD (intermodulation distortion) is typically only −25 dB (see **table 1**) — not very good if you want to be popular with other local operators.

Many high power tubes designed for linear service are costly or less efficient than the Class C-type tubes. This was a significant consideration, especially on EME, before the FCC changed the method of measuring power. That's all different now, and we're no longer penalized for using tubes with less efficiency. As a result, many less efficient tubes, especially those used in commercial television transmitters,

may now find their way into Amateur amplifiers after they're removed from commercial service during routine maintenance.

In summary, it would probably be best to design any new power amplifier with linear operation in mind. The small change in efficiency may easily be recovered by higher gain with less likelihood of generating key clicks. Using tubes other than those presently in wide use may also be advisable. **Table 1** and references 2, 3, and 4 should be consulted when choosing a suitable VHF/UHF transmitting tube.

configuration is important

One of the first considerations in designing a power amplifier is the tube type and quantity. Under the old FCC regulations, it was a common VHF/UHF practice to use two tubes in either a parallel or push-pull configuration. This was often done because efficient external anode tubes of moderate power (500 watts typical) were easily obtained, and usually at reasonable prices. Furthermore, many Amateur designs were already available in various publications. However, the new power limits are not easily achieved with the tubes that used to be popular — especially in linear operation.

If push-pull or parallel amplifiers are used, the tubes should be fairly well matched and from the same supplier. Extra metering and balancing circuitry are required to insure that both tubes properly share the load. If neutralization is used, this type of circuitry is doubly difficult. Therefore, it's better to use a single large tube that will do the entire job rather than multiple tubes that will share the load.

gain

The average amplifier gain in the VHF region is between 15 to 20 dB and 10 to 16 dB at UHF. As mentioned, gain is usually higher in linear operation than in Class C. A typical grounded-grid amplifier has 3 to 5 dB less gain but is usually easier to stabilize and match at the input. For example, on 2 meters a typical pair of grid-driven 4CX250Bs operating in Class C will typically deliver 400 to 600 watts of output with 10 watts of drive power, but only 200 to 300 watts on 70 cm (432 MHz). This will increase to 600 to 700 and 300 to 400 watts respectively in linear service. Hence, it should be obvious that an additional low-gain intermediate driver amplifier may be required if a typical 10-watt transverter is used to drive a power amplifier with a kilowatt or higher output capability.

input circuits

The input circuit for typical VHF/UHF amplifiers is one of two major types: either cathode or grid-driven. The cathode type usually requires only some simple "L" or "T" matching section (**fig. 1A**). Grid-driven amplifiers usually require a step-up or transformer type of input circuit with additional tuning and matching components (**figs. 1B** and **1C**).

If push-pull or parallel tube operation is used, circuitry should be designed to allow the drive to be tweaked for each tube (**fig. 1D**). If the input circuit doesn't have RF balancing capability, the input grid-biasing circuit should be designed to allow for independent adjustment of the DC bias voltage for each tube so that the load is shared equally between the tubes. The principal thing to remember when designing the input circuit of a power amplifier is that the configuration chosen primarily performs an impedance match between the input driver (usually a 50-ohm device) and the grid or cathode of the tube being driven. If the network is properly designed, the power amplifier driver will see a low VSWR (2:1 maximum).

stabilization

The screen grids of tetrode amplifiers must be properly bypassed. Most manufacturers and suppliers offer specially designed tube sockets with built-in bypass capacitors. These are highly recommended and, although usually expensive, will justify their cost with improved amplifier stability and efficiency. Some of the more modern tube socket designs (for

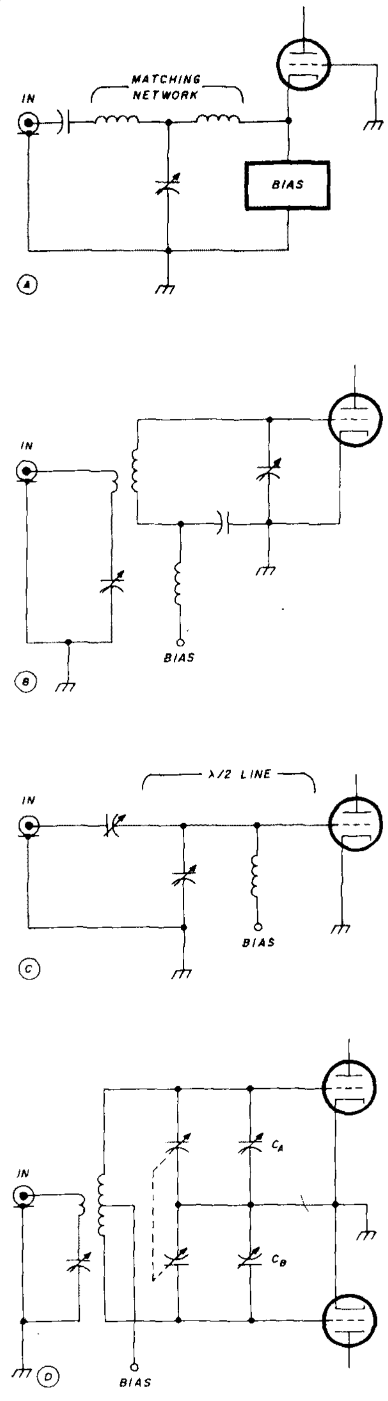


fig. 1. Typical input circuits used on high power VHF/UHF amplifiers. (*A*) grounded grid, cathode-driven, (*B*) single ended, grid-driven, (*C*) single ended 1/2 wavelength line, and (*D*) push-pull. C_A and C_B are balancing capacitors.

example, the EIMAC SK-620 and SK-630 types) also offer more input to output isolation and are highly recommended for new designs.

Neutralization may be required if the amplifier gain is high and/or if the inter-electrode capacitance between the grid and plate of the tube is high. A typical bridge neutralization network is shown in **fig. 2A**. Push-pull operation requires a slightly more elaborate scheme (**fig. 2B**).

Many VHFers prefer to avoid neutralization by using grounded grid circuitry, which doesn't usually require neutralization if the inductance of the grid (and screen grid, if applicable) bypass capacitor assembly can be kept very low. Additional information on neutralization and amplifier stabilization techniques is available.[1]

output circuits

There seem to be as many stories about VHF/UHF transmitter output circuits as there are types. The principal types are the π network, the resonant 1/4, 1/2, or 3/4-wavelength tank, push-pull, the rectangular cavity, and variations on the latter.

On 6 meters the π network is still quite popular because it's versatile, matches a wide range of impedances, and has some inherent harmonic reduction capability (**fig. 3A**). The main problem in using this technique is that the minimum loaded Q in the tank circuit is determined by the output capacitance of the tube in parallel with the π network input capacitor, C1. At HF this is typically 25 to 250 pF of capacitance, but only 8 to 15 pF at 6 meters, the output capacitance of the typical tubes being used!

If a high (>15) loaded Q is used, the input tuning capacitor, C1, should be kept to a minimum value. This requirement can best be met by using a flapper-type capacitor. Because of their inherent inductance, vacuum variables are not recommended; standard transmitting variables usually have too high a minimum value of capacitance. Some designers have circumvented this problem by choosing a loaded Q based on the tube's output capacitance and tuning the inductor, L1, with a shorted turn loop.[5,6]

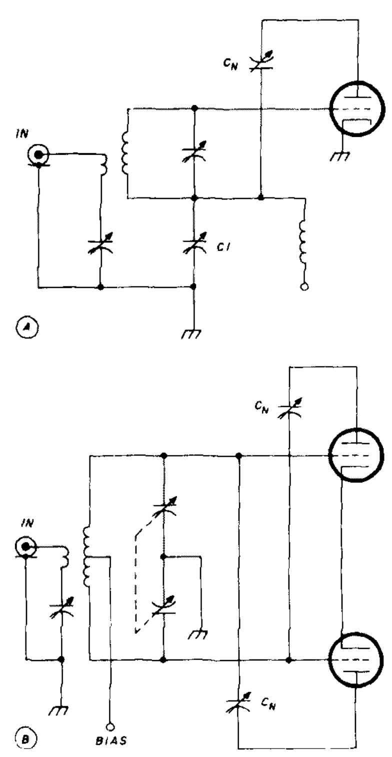


fig. 2. Typical VHF/UHF neutralization techniques. (*A*) single-ended "Bruene" circuit.[1] C1 should ideally be equal to the tube's input capacitance. C_N balances out the tube's feedback capacitance. (*B*) push-pull version.

On 2 meters through 70 cm, most Amateur amplifier designs typically use tank circuits employing either a 1/4, 1/2, or 3/4-wavelength coax, microstrip or stripline (**figs. 3B** through **3E**). Sutherland points out that the 1/4-wavelength line is preferred because it exhibits the greatest bandwidth, and further recommends that the plate line impedance should be between one and two times the capacitive reactance (plus strays) of the tube in use.[7]

If the output capacitance of the tube is too high at the frequency of operation, a 1/4-wavelength line may be impractical because of its very short length. Then a 1/2- or 3/4-wavelength line would be usable. One advantage of the 1/2-wavelength line is that it need not be fed with the tube on one end, but may instead have the tube at the center, as shown in **fig. 3D**.[8,9,10] In this configuration, the tube capacitance is effectively divided between each side of the line. This also makes tuning and loading easier. Another advantage of the balanced 1/2-wavelength line is that the current through the tube is more evenly distributed.

Because they offer a quick way to obtain higher power with smaller tubes,[11,12] parallel amplifiers are also quite popular. However, they are often difficult to balance and are really no more efficient than a push-pull type amplifier with proper balance and output coupling.

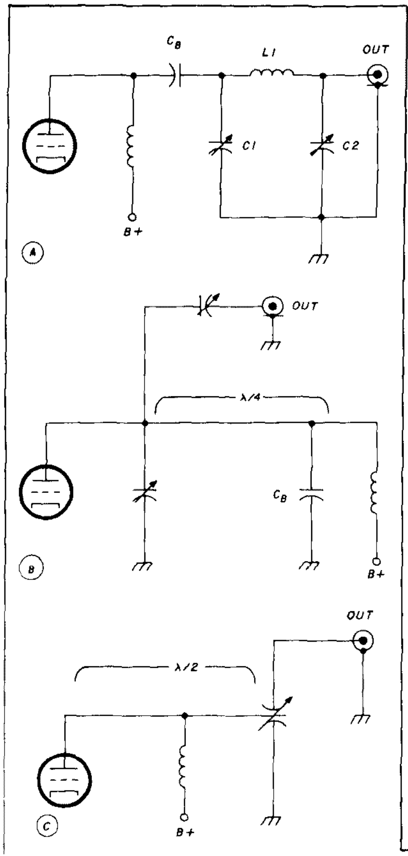
Push-pull circuitry (see **fig. 3F**) used to be very popular, especially with the 500-watt ceramic tetrodes;[13-17] it is easier to balance because each tube is fed separately. In addition, the symmetrical output configuration inherently suppresses the second harmonic.

One factor often ignored is the current flowing through the tube itself. With the typical single-ended microstrip line approach, there is often a concentration of current through one side of the tube. This can result in decreased tube life. Consequently, at UHF and microwave frequencies, the recommended configuration is the rectangular cavity (**fig. 3G**). This configuration is particularly useful at 70 cm and above.[18] Multiple tubes can be easily paralleled for higher output power.[19] The layout of the rectangular cavity usually assures that current is uniformly distributed throughout the tube.

Output tank circuits — of which there are many — are usually chosen according to the frequency of interest, the tube type used, and the number of tubes used. In a symmetrical output tank, the use of a single tube that can deliver the required output power is probably the best choice.

output coupling

As previously mentioned, the π network, sometimes used on 6 meters, is very easy to couple to the load (**fig. 3A**).



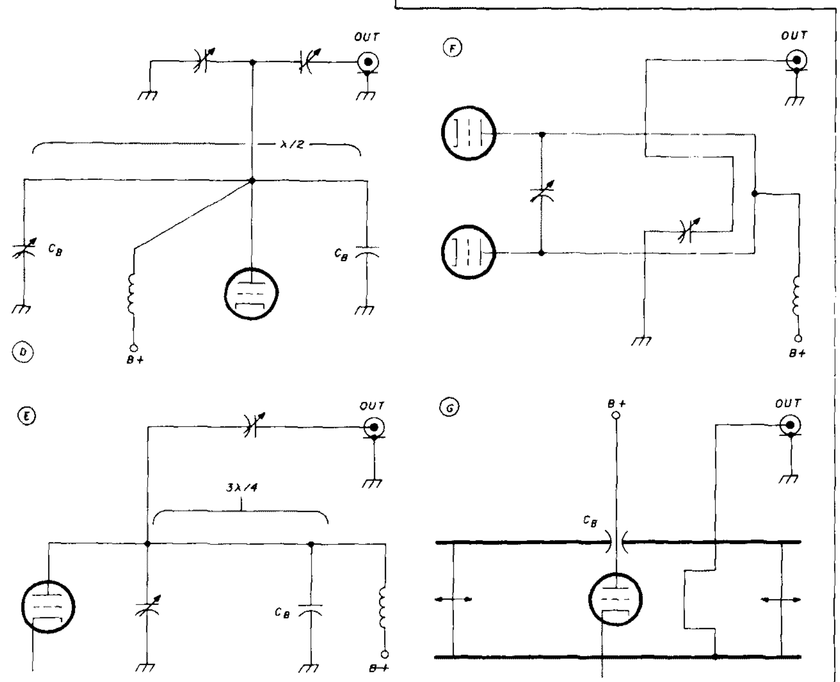


fig. 3. Typical output circuits. C_B is either a bypass or DC blocking capacitor. (*A*) π network, (*B*) 1/4 wavelength, (*C*) 1/2 wavelength, (*D*) 1/2 wavelength center-loaded, (*E*) 3/4 wavelength, (*F*) push-pull, and (*G*) rectangular cavity.

With the various types of tank circuits common in VHF/UHF amplifiers, output is often coupled by either an inductive link or by a series capacitor. While the former(**fig. 4A**) used to be popular, it has to be carefully designed and coupled to the tank circuit; this problem can be somewhat relieved by adding a series resonating capacitor in the loop (**fig. 4B**). If a loop Q of 2 to 5 is chosen, the capacitor will essentially operate as a loading capacitor, and the placement of the loop is not as critical. The resonant loop method is preferred because it reduces harmonic output.

Simple to design, use, and tune, the capacitance output coupling method shown in **fig. 4C** has become more popular for use in single-ended VHF/UHF amplifiers. But it can lead to problems, especially if the capacitance is increased beyond optimum, thereby causing overcoupling to the tank circuit. If this is done, harmonic output increases rapidly. Always use the least amount of coupling possible while maintaining acceptable output efficiency.

Let us not forget the push-pull configuration (**fig. 3F**). This is basically a balanced configuration, although its output network is frequently treated as unbalanced with the use of a series loop and capacitor, as just mentioned above. In fact, *some* users have claimed lower efficiency than expected. Efficiency can reportedly be increased by using capacitance output coupling in a balanced arrangement through a 4:1 half-wave balun.[20]

The rectangular cavity often uses a probe which is either a loop or a rod properly shunted across the cavity (**fig. 3G**). Once the proper location is found, tuning and loading can be accomplished by varying the spacing between the tube and the cavity walls, as shown in reference 19.

Finally, with the present-day designs, the proliferation of high power, and a congested VHF/UHF communications spectrum, we must pay more attention to out-of-band radiation and especially harmonics. Although a low-pass filter may reduce output power slightly. FCC regulations must be complied with. Remember also that if your amplifier has too much harmonic output power, it will appear to increase VSWR and yield false readings. In this regard, the push-pull configuration is preferred. Many harmonic suppression techniques are used, *but the simplest is probably* a shorted 1/4-wavelength coax stub shunted across the amplifier output connector.

summary

In this month's column I've only scratched the surface of VHF/UHF power amplifier design, emphasizing the new FCC output power regulations. Next month's column will explore the subject in greater depth, with emphasis on practical applications and construction. In the meantime, I'd suggest a review of the references listed — especially reference 1, because it deals with most of the problems prevalent in high power amplifiers and of-

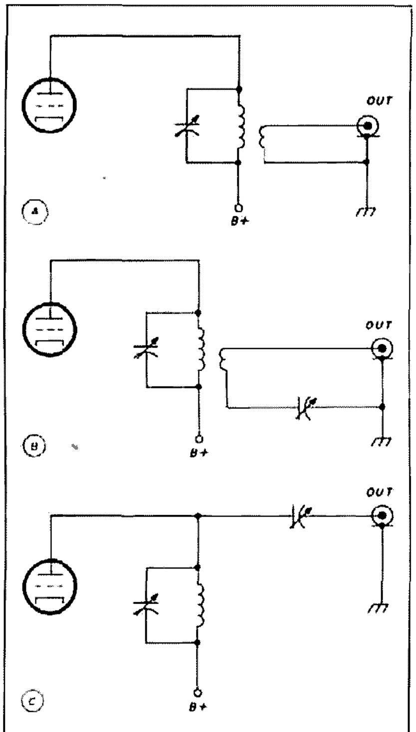


fig. 4. Typical output coupling techniques. (*A*) link coupling, (*B*) resonant link, and (*C*) capacitance probe.

fers many suggestions on solving them.

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January VHF/UHF events

January 3:	*0024 UTC, predicted peak of Quadrantid meteor shower*
January 12:	*EME perigee*
January 12-13:	*ARRL VHF Sweepstakes Contest*

ham radio

short circuit

VHF/UHF world

In W1JR's column, "VHF/UHF world: high dynamic range receivers," (November, page 97) the source of the NE41632B transistor and the balun core was correctly identified as PROTO-FAB. Since publication of the November issue, however, PROTO-FAB has changed its name to *PROTO-PARTS*.

PROTO-PARTS regrets that no telephone inquiries can be accepted. Mail inquiries, however, are welcome; please include SASE. Inquiries and orders should be addressed to PROTO-PARTS, 74 Wedgemere Drive, Lowell, Massachusetts 01852.

In the November column, certain paragraphs on page 100 were transposed. For a corrected copy of that page, send an SASE to *ham radio*, Greenville, New Hampshire 03048.

In **figs. 5** and **6** of W1JR's December column, a ground should be connected to the 5-volt, 3-terminal regulator.

product

REVIEW

frequency counter kit

The handheld frequency counter designed by Roger Ray, (as published in the UK's *Radio and Electronics World* in November, 1982, and imported and sold by RADIOKIT) is a neat little project that shouldn't take up too much time. A valuable piece of test equipment for your ham shack, it runs off an inexpensive 9-volt battery, is quite light and easy to carry around, covers 20 Hz to 150 MHz, and has a five digit LCD display with a resolution of 1 Hz to 10 kHz depending upon range selected.

design

Ray went to great lengths to design this meter with simplicity in mind. An FC-177 LCD readout module, which also contains an OKI MSM-5527 frequency counter, is used for the display. Combining these two functions into one package greatly simplifies the construction of the counter. The FC-177 is designed to measure and display frequencies from 20 Hz to 3.999 MHz. To measure frequencies higher than 3.999 MHz, a divide-by 10 or a divide-by-100 prescaler is incorporated. To keep power consumption low, a count-and-hold technique is used when measuring frequencies above 3.999 MHz. In the LF range, 20 Hz - 10 kHz, the unit has a 50 mV sensitivity; in the MF range (10 kHz - 4 MHz) a 20 mV sensitivity; in the HF range (100 kHz - 40 MHz) a 20 mV sensitivity; in the VHF range (10 MHz - 150 MHz) a 100 mV sensitivity. Users will find that this unit, while not designed for laboratory precision, will give more than adequate readings for nearly all of their measurements.

circuit description

The input frequency is switched by the frequency range selector to one of four buffer stages. For measuring in the LF range, the signal is amplified, its frequency is multiplied by 100, and then used to control a VCO that is a part of a PLL circuit. The frequency displayed is that of the VCO with the decimal point properly positioned to account for the X100 factor. For example, an 800 Hz input signal would be changed to 80 kHz through the VCO, but counted and displayed as 800 by the FC-177.

Measuring in the MF range is within the FC-177's design. Signals are amplified and then directly fed to the display module. To measure in the HF range, signals are first amplified and then fed through a divide-by-10 prescaler (MSL-231RS) directly into the FC-177 for counting and display.

VHF range measurements are accomplished in a manner similar to that of HF signals, with the prescaler changed to divide-by-100.

As mentioned, this unit is designed for low current consumption. However, the MSL-231RS is designed with a current consumption of greater than 30 mA. To reduce current demand, a hold feature of the FC-177 is employed. Instead of constantly being counted, the incoming frequency is measured once every second. The designer calculates that this reduces consumption to below 15 mA, has little effect on accuracy, and adds several hours of battery life.

construction

The unit is built around a single-sided PC board and mounted to the enclosure by the four-way frequency switch. The FC-177 is mounted on the cover of the plastic equipment box. Parts placement is fairly straightforward, with connections to the FC-177 frequency counter module through short flexible wires. All IC's are mounted on sockets so they can be replaced with a minimum of effort in the unlikely event of failure.

Careful attention to the location of the display, four-way switch, momentary on-off switch and BNC input connector is a must if the unit is to function properly.

I'd estimate that overall time to build shouldn't be more than an evening or two, barring any unforeseen difficulties.

conclusion

At $74.95, it's really hard to beat this unit for ease of construction and usefulness in the ham shack. RADIOKIT has a number of other projects from various amateur radio publications. For a free catalog, contact RADIOKIT, Box 411H, Greenville, New Hampshire 03048.

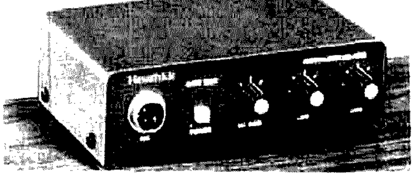
Circle #179 on Reader Service Card.

— **N1ACH**

NEW products

microphone equalizer

Heath's new HD-1986 Microlizer is designed to improve the quality of transmitted speech and provide a better match between microphone and transceiver. This battery-powered microphone



equalizer fits in series with a microphone and transceiver using a standard 4-pin microphone jack and 1/4-inch phono output jack. It has continuously variable frequency controls to provide a ± 12 dB (boost and cut) at 490 Hz and 2800 Hz. A gain control permits the user to increase or decrease the microphone signal fed to the transceiver for maximum efficiency and cleaner operation. The Microlizer can be bypassed to allow direct connection between microphone and transceiver by simply turning off the power switch.

For complete information and/or a copy of the current catalog, contact Heath Company, Department 150-405, Benton Harbor, Michigan 49022.

Circle #301 on Reader Service Card.

universal audio filter

Palomar Engineers has announced a new universal receiver audio filter. Model FL-4 — for SSB, CW, and RTTY — features switched capacitor filters. A 10-pole low-pass and an 8-pole high-pass can be moved anywhere in the 200-3500 Hz range to form a sharp bandpass filter at any frequency and of any bandwidth. A notch filter is also included.

It connects to the receiver phone jack and provides 2 watts of audio to drive a speaker. The on-off switch bypasses the filter when not in use. It operates from 15 VDC. The price is $139.95 plus $4 shipping. An optional 115-VAC adapter is available at $9.95.

For further information, contact Palomar Engineers, Box 455, Escondido, California 92025.

Circle #302 on Reader Service Card.

COR module

Hamtronics, Inc., has announced the COR-3, a new version of its popular COR module. Like the COR-2, the COR-3 has all the circuitry needed to control a transmitter and receiver to make a repeater, including an electronic relay to switch the transmitter on and off as a function of the receiver squelch, a tail timer, a time-out timer, an audio mixer, and a local speaker amplifier. The COR-3 also has a "courtesy beep" function, and an additional timer that allows the beep to be adjusted up to five seconds after the receiver squelch drops. Whenever a station using the repeater releases its microphone, a beep tone is heard after a short delay period. The beep indicates that the party has finished talking and the time-out timer is reset.

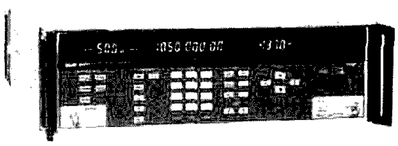
The price of the COR-3 kit is $58. For more information on this module and other transmitter, receiver, and control modules for building repeaters, contact Hamtronics, Inc., 65F Moul Road, Hilton, New York 14468-9535.

Circle #104 on Reader Service Card.

NEW products

signal generator

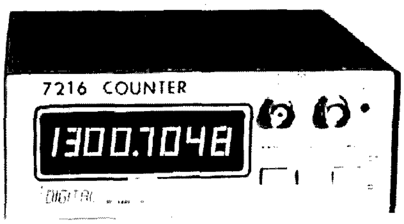
A programmable, general-purpose signal generator base-priced at $4500, is said to meet or exceed the quality and performance of units costing over $6000. The Fluke 6060A Synthesized Signal Generator accurately tests a wide variety of RF receivers, filters, amplifiers, and mixers. It covers a frequency range of 0.1 to 1050 MHz; (selectable with 10 Hz resolution) and has a switching speed less than 100 ms typical. Non-harmonic spurious products are less than 60 dBc, and harmonics are less than −30 dBc across the entire frequency range. Amplitude levels are selectable from −137 dBm to +13 dBm with 0.1 dB resolution.



For further information, contact John Fluke Manufacturing Co., Inc., P. O. Box C9090, Everett, Washington 98206.

1.3 GHz frequency counter

Digital Instruments Inc. (formerly David Electronics) of Tonawanda, New York, has announced its frequency counter (#7216). The new counter has a range of 10 Hz to 1.3 GHz and a gate time of 100 MHz 0.1 and 1.0 second as well as 1.3 GHz 0.16 and 1.6 seconds. Its display consists of eight 0.04-inch LEDs with an automatic



decimal point. The prescaler and built-in gate light all fit neatly into the small 5-1/2 x 6 x 2-inch all-metal case. Its power requirements are 105-125 volts 50/60 MHz at 3 watts with a safe input of 120 volts RMS to 10 MHz and 2 volts RMS above 50 MHz. The price is $249.95.

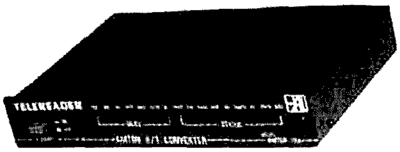
For additional information contact Digital Instruments, 636 Sheridan Drive, Tonawanda, New York 14450.

Circle #136 on Reader Service Card.

NEW products

AMTOR code converter

The AMTOR-10A unit from HAL allows transmission and reception of radio teleprinter signals with the added feature of error correction. The AMTOR-10A is a code and speed (baud rate) converter unit that, when used with



the CWR6850, allows 7-unit AMTOR code communications at 100 baud while retaining communications capability with the standard 5-unit BAUDOT code from 20 to 100 baud and the 8-unit ASCII code from 75 to 200 baud. The AMTOR-10A is designed to interlace easily with the CWR6850 while retaining all the standard features of the unit, including CW operational capability. The price of the AMTOR-10A code converter is $395.00.

For further information contact HAL Communications Corporation, P.O. Box 365, Urbana, Illinois 61801.

Circle #303 on Reader Service Card.

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REALISTIC 5-band solid state communications receiver, short wave and domestic, with speaker, efficient plastic dipole room antenna, with calibration charts and manual. Four years old. Used very little and has excellent world wide reception. $70 or best offer. Paul S. Bassett, Jr., RFD #2, Sutton Center, MA 01527.

HEATHKIT VF7401 — 2 meter FM digital scanning transceiver. Checked out by Heath. Asking $225. Call (217) 344-4265. Tim Lazaro, W9OEN, 203 S. Birch St., Urbana, IL 61801

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TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details. W8UUR, 1025 Wildwood Road, Quincy, MI 49082.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

DIGITAL AUTOMATIC DISPLAYS for Collins, Drake, TS-520S (functional DG-5 replacement) and Swan 350C through 700CX. No installation. Plugs into existing jacks and tube sockets (using extenders). FT-101's and TS-520 models are easily installed. No drilling. Six red 1/2" digits in a 5" by 1-1/2" by 9" deep cabinet! Write for information. Grand Systems, PO Box 2171, Blaine, WA 98230. (604) 530-4551.

REPAIR, ALIGNMENT, calibration. Collins written estimates $25; non Collins $50. K1MAN. (207) 495-2215.

ATLAS 350XL owners group. Send QSL card with s/n your rig. Know anyone who repairs them? Have any technical information to share? Any questions? Rod, N5NM, Box 2169, Santa Fe, NM 87504.

CHASSIS and cabinet kits. SASE K3IWK.

SCHEMATICS: Radio receivers 1920/60's. Send name brand, model, SASE, Scaramella. P.O. Box 1, Woonsocket, RI 02895-0001. (602) 897-2534.

AMATEUR RADIO's newspaper — Worldradio. Latest info. One year subscription (12 issues) only $10. Worldradio, 2120-B 28th St., Sacramento, CA 95818.

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HAM RADIO Magazine collection: Bound volumes 1972, 1973, 1974 and 1975. In HR binders 1976, 1977, 1978 and 1979. $120 for lot plus UPS shipping. US only. W4UCH, Box 1065, Chautauqua, New York 14722 (716) 753-2654.

WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093. (312) 251-8901.

$$$$ SUPER SAVINGS on electronics parts, components, supplies, and computer accessories. Free 40-page catalog for SASE. Get on our mailing list. BCD Electro, PO Box 830119, Richardson, TX 75083. Or call (214) 690-1102.

WANTED: Old microphones, remote mixers other misc related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

WANTED: Used microwave equipment. D. Mitchell, K8URL, 1 Cider Mill Lane, Upton, MA 01568. (617) 529-4638.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

VHF, UHF, ATC kits high quality professional kits from Wood & Douglas, England. Designs for RX's, TX's, Xcvr's, preamps. Catalog. Lance Lyman, K8IXZ, Tactical Electronics Corp., PO Box 1743, Melbourne, FL 32902.

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skyrider Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

VERY in-ter-est-ing! Next 4 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60189.

GENERAL RADIO 650A impedance bridge $75. 650-P1 oscillator amplifier $75. WA3WHR, 4109 Queen Mary Drive, Olney, MD 20832 (301) 774-9131.

SCHOTTKY mixer diodes HP-2800, 8/$1 + 25¢ postage. Koempel, Box 335, Keansburg, NJ 07734.

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COMING EVENTS

Activities — "Places to go . . ."

INDIANA: South Bend Hamfest Swap & Shop, January 6, first Sunday after New Year's Day at Century Center downtown on US 33 Oneway North between St. Joseph Bank Building and river. Industrial history Museum in same building. Carpeted half acre room. Open tables $1 per ft. Four lane highways to door from all directions. Talk-in freq: 52-52, 99-39, 93-33, 78½18, 69-09, 145.29.

AFCEA Amateur Radio Luncheon sponsored by the Armed Forces Communications and Electronics Association at their 6th Western Conference held at the Disneyland Hotel in Anaheim, California, 29 through 31 January 1985. Admission free to all registrants. There is a nominal charge for luncheon tickets. Following the meal, a panel of distinguished leaders will participate in a forum on "Amateur Radio Support for the National Communications System". For additional information: John W. Browning, W6SP, 6202 Lochvale, Palos Verdes, CA 90274. (213) 544-2543.

ARIZONA: The Eastern AZ Amateur Radio Society will hold Amateur Radio license exams, Saturday, January 26, 9 AM to 3 PM, National Guard Armory, 4001 First Avenue, Safford. Prior registration deadline January 21. Send completed form 610, copy of license and $4.00 registration fee to EAARS, PO Box 402, Thatcher, AZ 85552. For information: Richard, N7DZH (602) 428-6560 7 AM to 3:30 PM.

MASSACHUSETTS: The Mount Tom Amateur Repeater Association will host its first annual indoor Flea Market, March 3, 9 AM to 4 PM, Knights of Columbus Hall, Elder Council 69, Granby Road, Chicopee. General admission $1.00. Kids and spouse free. Tables $8.00 door; $7.00 advance. Set up 8-9 AM. Food and drink. Contact Mickey Yale, N1CDR, 6 Laurel Terrace, Westfield, MA 01085. (413) 562-1027.

NEW YORK: Yonkers Electronics Auction, Sunday, January 27, 9 AM to 3 PM, sponsored by the Yonkers Amateur Radio Club. All indoors at Lemko Hall, 556 Yonkers Avenue. Admission $3.00. Children under 8 free. Inspection from 9-10 AM. Club commission on successful sales only 10% first $100; 5% remainder. Hams computer and electronic wizards, CBers bring equipment (new and used) you want to auction off. Unlimited free coffee all day. Talk in: 146.265T/146.865R, 52 direct. 440.150T/445.150R. For information: YARC, 53 Hayward Street, Yonkers, NY 10704. (914) 969-1053.

LOUISIANA: The Southeastern LA University ARC (SLUARC) and the Southeast LA ARC (SELARC) are jointly sponsoring a Hamfest, Saturday, January 19, 9 AM to 3 PM at the old men's gym on the Southeastern LA University Campus. Free admission.

THE VOLUNTEER EXAMINERS of the Grand Rapids Amateur Radio Association in cooperation with ARRL will conduct Amateur Radio exams in Grand Rapids, Michigan on the following dates: Friday, February 15, Friday, June 21, Friday, October 18, 1985 and Friday, February 21, 1986. Mail FCC Form 610, check/MO for $4.00 made out to ARRL/VEC to: ARRL/FCC Amateur Testing, c/o Mike Bottema, K8EX, 930 — 92nd Street, SE, Byron Center, MI 49315.

Operating Events — "Things to do . . ."

JANUARY 26: West Virginia QSO Party, 1700Z Jan. 26 to 1700Z Jan. 27. Single operator only. Exchange signal report, serial number and QTH (county for WV stations; state or country for others). Mail logs by Feb. 11 (include large SASE for results) to K8DS, PO Box 1694, Charleston, WV 25326.

YL-OM CONTEST. Phone: Start St. February 9, 1800 UTC ends Sun. Feb. 10 1800 UTC. CW start Sat. Feb. 23 at 1800 UTC ends Sun. Feb. 24 at 1800 UTC. All licensed men and women operators throughout the world are invited to participate. Exchange station worked, QSO number, RS or RST, ARRL section or country. Entries in log must also show time, band, date and transmitter power. Logs must show claimed score and be postmarked by March 15, 1985 and received no later than April 5, 1985. Please send logs to: Marty Silver, NY4H, 3118 Eton Road, Raleigh, NC 27608, USA.

VT QSO PARTY 1985 0001Z February 2 to 2400Z February 3. Exchange VT stations send RS(T) and county (CW) two-letter county designators. Other stations send RS(T) and state, province or ARRL country. Send SASE now for official score and log sheets. Send logs/facsimiles, name, address, county (VT), nlt March 1, 1985 to: D. Nevin, KK1U, W. Hill, Northfield, VT 05663.

1985 NEW HAMPSHIRE QST PARTY, 1900Z February 2 to 0700Z February 3 and 1400Z February 3 to 0200Z February 4. Work stations once per band and mode. NH to NH QSO's allowed. Exchange signal report and QTH (county for NH stations, state, VE province or DXCC country for others). Logs must be postmarked by March 15. Include large SASE for results. Mail logs to: Great Bay Radio Assoc., PO Box 911, Dover, NH 03820.

WEST COAST 160 BULLETIN SSB CONTEST, February 9 0000 GMT to February 10 2359 GMT. Single operator only. Exchange RST, QTH. Subscribers/non subscribers. Score 10 points per QSO. Multipliers: states, VE Prov. Country. Log info: date, time, rst, QTH. Send logs to: R. Koziomkowski, 5 Watson Drive, Portsmouth, RI 02871. Logs must be postmarked before March 31, 1985.

THE GUERRI REPORT

Ernie Guerri
W6MGI

spectrum utilization: a challenge to technology

In almost every segment of the radio spectrum we're faced with increasing demands for frequency assignments, greater bandwidths, and increased radiated power (range/reliability). Some specific segments groan under the burden: the AM and FM broadcast bands, the 40 and 20-meters Amateur bands, VHF and UHF business radio, and the 2 GHz microwave area. Spectrum conservation and improved utilization techniques have become a high priority for practically every user of the airwaves.

Until now the methods used to squeeze more signals into the same tight space have been relatively simple: use SSB, reduce FM deviation, improve antenna directivity, and so on. But more sophisticated techniques *will* be needed if we're to make more efficient use of the spectrum.

Fortunately, in 1948 a thoughtful scientist — Claude Shannon at Bell Laboratories — developed a theory of information transmission that showed the relationship between speed, bandwidth, and time in a usable mathematical form. Dr. Shannon's work made possible the coding concepts that permit more effective utilization of spectrum space. There are several ways in which RF signals can share the same "space":

- time sharing
- frequency sharing
- different antenna polarizations
- coding

The goal of each of these techniques is to yield a signal-to-noise ratio that conveys information (a change of data) usable to the data sink — which is frequently a person, but could also be an unattended data terminal. Indeed, time, frequency, and polarity diversity are all forms of coding.

Modern computer technology gives us options with respect to code complexity, efficiency, and speed that were simply not available when our present modes of communications were being developed. The basic objective of data coding for spectrum efficiency is to omit as much data as possible while still conveying relevant information.

One of the more successful techniques for bandwidth reduction is being used by some computer manufacturers to permit very high resolution graphics on conventional RGB displays. This process is called bit-plane encoding. In this process a signal of 2^n possible amplitudes is transformed into n signals, each of which has only two amplitudes. The 2^n amplitudes subsequently consist of an n-bit binary word at the output of an appropriate quantizer. This process has demonstrated that it can reduce by six times the bandwidth needed to transmit high quality, full motion TV images. Manufacturers are now developing dedicated digital signal processing chips to perform the necessary bandwidth compression and S/N ratio enhancement functions.

If we remember that information requires a *change* of data, then even more bandwidth reduction is possible by further reducing, or eliminating, redundant data. AT&T adopted this technique, — called "conditional replenishment" — to make possible the "picture-phone." In this approach, a TV frame is stored in a memory and compared against a subsequent frame. Only the parts that are different are transmitted. It was found that a bandwidth of less than 100 kHz could convey an acceptable moving picture using this method. If variable persistence and digital background refresh are available at the receiving end, the data needs to be sent even less frequently. Think of how much redundant data is conveyed in the Amateur bands — background noise, non-linear distortion, excess power when band conditions are good, and so on. If Amateur SSB/FM rigs just had some digital storage and "variable persistence" audio output stages . . .!

Some TV receivers utilizing these techniques may be available by next year, but the real challenge remains a general commitment by the electronic industry to more frequent implementation of modern techniques.

Since much of the processing needed to effect significant bandwidth reduction is very complex, most Amateurs will have to wait until the chips are readily available before they'll be able to actually use these techniques in hardware.

Even more efficient than these techniques, but still years away from Amateur implementation, are mutually adaptive data links. This approach enables both ends of the link to regularly adjust their own performance to accommodate the predetermined acceptable data quality.

Although this column is reserved for discussions of technological trends

with implications for Amateur Radio, the need for better spectrum management is so urgent that I can't help but offer a few comments about ways in which each of us can help assure better use of our present bands:

• Use the minimum power necessary. If conditions are good, settle for S9 on the other end. Remember that the mike gain is as useful as the volume control.

• Use filters — keep out-of-band harmonics and spurious responses to a minimum. Upgrade the internal filters (crystal/mechanical) in your rig to units with steeper skirts if they're available. Audio filters and response shaping can be useful in both the microphone and speaker circuits at your station.

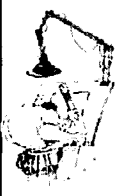
• When possible, use directive antennas and keep the main lobe on the station you're working. Never mind proving that you can work Lonely Island off the back of your beam.

• Operate your rig within its limits. If you run a rig rated at 1 kW, keep it at 1 kW. Pushing it to 1500 watts will only generate distortion, won't help you at the receiving end, and will probably cause the stations on either side of your frequency to miss the opportunity altogether.

Someday the electronic capabilities I've discussed will compensate for the effects of individual operating habits on the spectrum. Until that time comes, solving the problem of effective spectrum utilization will be up to us.

ham radio

WANTED: COVER ART



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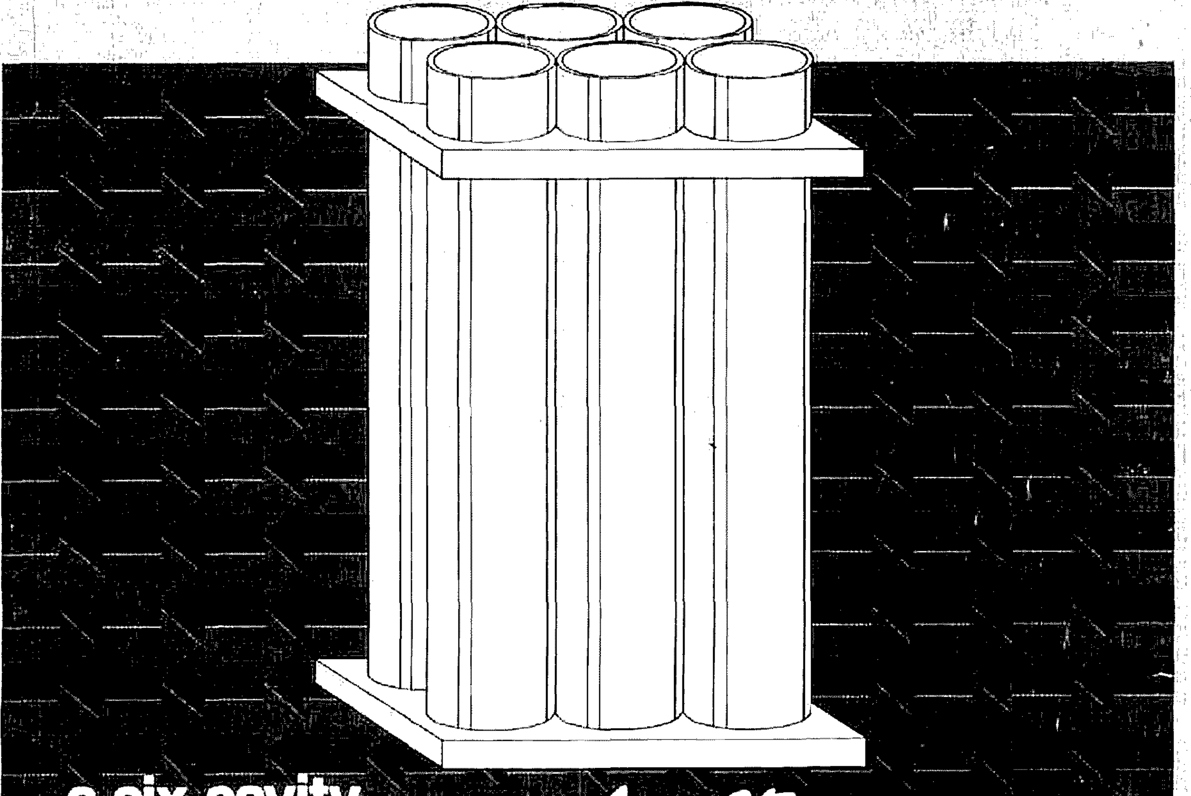

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FEBRUARY 1985

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contents

REFLECTIONS

one million years of experience

If you saw a resume that showed one million years of experience, I'm sure you'd be impressed. One million, after all, is a very large number. But that, roughly, is the amount of experience that we Radio Amateurs have to draw on. In fact, if you add all our years of experience together, it's probably two or three times that number.

More and more often it seems that we're being told that Amateur Radio is dying, that our numbers are diminishing day by day. If you examine the figures — the inanimate statistics — it would, sadly, appear to be so.

The key word here is *inanimate*. Yes, the numbers are inanimate. But we are not. Just thinking of what hams have been able to accomplish, and continue to accomplish, over such a short period of time, is mind-boggling. Perhaps what we need to do is stand back and look at the future of Amateur Radio from a broader perspective.

Please understand that I'm not for one minute discounting the facts or the importance of the statements made by others who share our common interest in the preservation of our hobby. Call me a perennial optimist . . . it's just that when I look at what hams are doing in this country and abroad, I find myself feeling that we do have the capability to bring about the needed changes — but only if we are convinced of the urgency of the situation.

Before putting pen to paper I scratched my head to come up with positive suggestions for actions that would reverse the "doomsday" trend suggested by the declining numbers. ("Surely editors must have greater insight into solving the problems of their own field," I thought. "Given time," I supposed, "I'll contribute a suggestion or two that can be put into action.") But then I thought about you, the half-million hams who have their own opinions about what's happening, and about what can and should be done to strengthen and improve our hobby. What an unbelievable resource!

Maybe *ham radio's* first contribution can be to act as a clearinghouse for information. It would be our pleasure — nay, our responsibility — to help in collecting, sorting, and sharing your written suggestions, ideas, and insights about how to encourage the growth and expansion of Amateur Radio.

Quick! While the thought is still fresh, jot it down on a blank QSL card, post card, or letter addressed to me at *ham radio*, Greenville, New Hampshire 03048. I promise that I'll read each and every one.

Even the whisper of an idea can evolve into a plan and finally into action. "Sure," you can say, "this is all very general, but what can I add?"

That's simple. What are *you* interested in? What aspect of Amateur Radio do you want to preserve and see grow? What do you like about our hobby? Dislike? What would you like to see change? And how would you change it?

Ah, but who has the time? And what good will it do? These days, none of us has the time. We're all so busy, busier than we've ever been. But my answer to this is simply, "Nothing ventured, nothing gained." I'm willing to devote many hours to reading your responses — which may take no more than a minute or two to write.

Make me very popular at the Greenville Post Office. Send those suggestions in today, tomorrow. Keep them coming.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

BURBANK, ILLINOIS' ANTENNA ORDINANCE HAS BEEN EFFECTIVELY OVERTURNED under the terms of a Consent Decree entered in U.S. District Court for the Northern District of Illinois on November 30. Burbank's highly restrictive ordinance had put a one-year moratorium on new antennas, which it limited to 35 feet, required insurance plus a $50,000 bond and annual inspection fees, and authorized a fine for interference to home entertainment devices.

The Successful Two-Year Court Fight Was Waged by attorney Jim O'Connell, W9WU, on behalf of WA9EKA and 58 other Burbank Amateur and CB operators. Under the terms of the settlement Burbank agreed to grandfather all existing antennas, promptly issue permits ($15 maximum fee) for new towers up to 65 feet (exclusive of any antenna!), and permit roof mounting up to 12 feet above a building without permit. In addition Burbank is under court order to repeal both the offending antenna ordinance and any other city ordinances or codes in conflict with the terms of agreement. December 19 the Burbank City Council unanimously passed the new ordinance required by the agreement.

Estimated Costs Of This Important Battle Are Over $25,000 for the Amateur community alone, not including the tremendous investment in participants' time. The cost to Burbank taxpayers is not known. Though Burbank does represent an important victory for those Amateurs involved, the fact that it was by Consent Decree (which means in essence Burbank gave up rather than continuing to fight) somewhat diminishes its value as a precedent.

PRB-1, The ARRL's Attempt To Get FCC's "Official Sanction" for Amateur Radio against local restrictions, received strong support from several non-Amateur sources before the Comment period closed in late December. The American Red Cross and a number of communities and county emergency organizations have all joined in supporting the principals of PRB-1. However, it appears unlikely that the League petition will see any Commission response in the near future, quite possibly not until mid-1985.

20 KHZ CHANNELS ON 2 METERS' TOP HALF COMES TO THE MIDWEST, following overwhelming approval of the change by the Michigan Area Repeater Council at its December meeting. The timetable requires their frequency coordinator to come up with a comprehensive plan for changing existing repeaters' frequencies by next June. The actual changeover has been set to take place during May, 1986.

To Accomplish The Switch Will Require Moving All present "split" (15 kHz) systems plus half those in present 30 kHz slots. Unfortunately, 20 kHz channels provide only 99 slots in 2 MHz vs the 132 available in the present scheme. However, 15 kHz spacing has never been entirely satisfactory, while those areas that have already made the change to 20 kHz report they now have few if any problems with adjacent channel interference.

The Shift To 20 kHz Began In The Pacific Northwest, starting with British Columbia and Washington, then Oregon, Idaho, Montana, Utah, Arizona, and now Michigan. In addition, it appears to have been mandated for Mexican Amateurs by their government, and Texas, Louisiana, Kansas, Nebraska, and Oklahoma are all reported seriously considering the change.

The Impact Will Fall Directly On Major Population Centers in adjacent states as well as Canada, so some sort of response to Michigan's action is expected soon. A meeting has been called by ARRL Great Lakes Director W4OYI for January 19 in Ft. Wayne to discuss the situation; how well attended it will be remains to be seen, since it's being held the day before the Midwest's biggest winter hamfest in Arlington Heights, Illinois, 185 miles away.

2240 AMATEUR EXAMINATION ELEMENTS WERE ADMINISTERED IN NOVEMBER by volunteer examiners, with an overall pass rate of 48%. Top Regional VEC was DeVry, whose VEs gave 298 elements wiht a 53% pass rate. Runners-up were GLAARG (Los Angeles), with 225; ARRL 4th District, 212; Central Alabama, 193; and Metroplex, with 185. Nationally ARRL's groups gave a total of 564 elements, and W5YI's 135. DeVry also led in number of exam sessions during the month, with 16. Of the 51 VECs in place, only 55% were active in November.

A Net Devoted To The Volunteer Exam Program Meets Every Sunday morning on 7280 kHz at 1700Z. Net Control is W9JUG, who heads the VEC program at DeVry, and though the net is principally for coordination of the ninth call area VE operations any Amateur who is interested in the volunteer exam program is invited to join in.

A NEW BAND PLAN FOR THE 13 CM BAND, WHICH RECENTLY LOST 80 MHz to telemetry by FCC action, is going to have to be devised by the ARRL's VUAC. The two segments that remain, 2300-2310 and 2390-2450 MHz, will have to be reallocated to accommodate such diverse users as moonbounce, fast scan TV, and the Amateur Satellite Service (which has been authorized, though not exclusively, 2400-2450 MHz by WARC 79). Comments and suggestions should go to VUAC Chairman Dick Jansson, WD4FAB, or to Mark Wilson, AA2Z, at the ARRL.

OSCAR 10 IS ON A REDUCED OPERATING SCHEDULE for at least the next few months, to reduce battery drain during a period of partial eclipse of its solar panels. Check the Tuesday night or Sunday AMSAT nets for current times and modes.

Amateur Satellite Orbital Predictions For 1985 Are Available again from Project OSCAR. Their 1985 orbital calendar covers all four Russian Mode A transponders, RS5, 6, 7, and 8, plus all necessary data to determine the apogee of each OSCAR 10 orbit. Minimum donation for U.S. and Canada users is $10 (it's $12 overseas), to Project OSCAR, Inc., Box 1136, Los Gatos, California 94022. Please include a self addressed mailing label, too.

comments

half-wave sloper

Dear HR:

The response to my ham note, "80-meter Half-wave Sloper Uses Reflector," (October, 1984, page 48), has been excellent. One interesting note: *ham radio* arrived at our QTH the Friday before the Boxboro (Massachusetts) hamfest. We spent the weekend at the hamfest. Monday morning I called CQ DX 80 (grayline) and got Graham, ZL3MZ. Into the contact, Jim, KF4HK, broke in, asking for a report from the ZL. (He got 5-9.) He then informed me that he had put up two of my slopers over the weekend, worked a ZS the night before, and wanted to say "Hi." A quick check — with others helping — showed a 3 s-unit (18-dB) front-to-back between his two slopers.

Bruce A. Clark, KO1F
Belfast, Maine

J-pole or Zepp?

Dear HR:

The J-pole antenna described on page 43 of the July issue of *ham radio* (see "All-metal 2-meter J-pole Antenna," by Michael Hood, KD8JB) is not the magical 5/8-wave radiator that the author describes. The radiating portion is only that which extends above the 19-inch 1/4-wave matching transformer. The radiating element is therefore 38 inches, which is a 1/2-wave end-fed "Zepp."

Remember, the 1/4-wave transmission line inverts the high impedance to a low one. A short at one end insures a low impedance.

Slide the feedline away from the shorted end until a match is found.

The reason "convention dictates that the antenna point upward" is so it won't interact with the feedline and distort the radiation pattern.

For more information, check your antenna handbook under the index title "Zepp or End Feed."

Richard Ociepka, K1WWT
Augusta, Maine

The intent of my article was not to design a J-pole antenna, but rather to adapt a number of approaches to building a J-pole antenna for my own use. Along the way, I mentioned why I felt the J-pole antenna was a viable alternative to using a 1/4-wave radiator which had been the main 2-meter antenna at KD8JB to this point. I did not intend for anyone to assume this antenna was a magical 5/8-wave radiator.

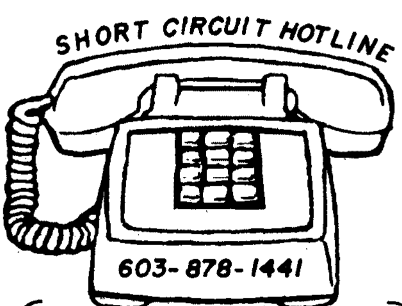
K1WWT is indeed correct in that the J-pole is an end-fed antenna and could be compared to the Zepp, since it is also end-fed, but I wouldn't go so far as to say the J-pole actually is an end-fed two-meter Zepp antenna. While it resembles the Zepp schematically, its appearance does not resemble the Zepp's any more than a Vee, inverted Vee, driven element of a Yagi, or other form of 1/2-wave center-fed radiator resembles the basic dipole. These derive their names from their shapes. If I offended K1WTT by not calling the J-pole an end-fed Zepp, my apologies — but I'm still going to call it a J-pole, because that's what it looks like, which was why someone (not me) selected that name in the first place.

The Zepp is a 3/4-wave antenna in its true form operating against a counterpoise of 1/4-wavelength. The J-pole is tapped by the feedline at roughly the 5/8's point on the radiator — hence my calling it a 5/8-wave antenna. (I'm not the only one calling it a 5/8-wave antenna, either.) We can argue this point forever, but I don't think we'll get any further than we are.

As far as "convention dictating the antenna be mounted upwards," good engineering practice dictates that the feedline of any antenna be brought away from the antenna perpendicularly, or at right angles (your choice — same result) for at least 1/2-wavelength to keep coax/antenna interaction to a minimum. If that is indeed the case, then the installation off the side of the tower as I mentioned in the article is correct, and actually better than if you were to run the cable straight down and away from the antenna, as is most commonly done when a vertical antenna is mounted on top of a tower. I suppose there would be times when interaction would occur regardless of how the antenna was installed, but my experience to date has shown that no adverse effects have manifested themselves by pointing the radiator in the downward direction. It's difficult to argue with success.

While not perfect by any stretch of the imagination, the J-pole as I built it works as I had intended it to work for my purposes. In addition, I felt that hams who put in a 40-hour week do not want to spend their free time with their noses in antenna engineering books trying to build the perfect antenna for their two-meter base stations. They want it quick, and they want it to work. I'd thought I'd covered all those bases.

Michael P. Hood, KD8JB
Grand Rapids, Michigan



Building a current *ham radio* project? Call the Short Circuit Hotline any time between 9 AM and Noon, or 1 to 3 PM — Eastern time — *before* you begin construction. We'll let you know of any changes or corrections that should be made to the article describing your project.

(See "Publisher's Log," April, 1984, page 6, for details.)

a home-brewed
six-cavity duplexer

Achieve 95 dB of isolation, using inexpensive parts

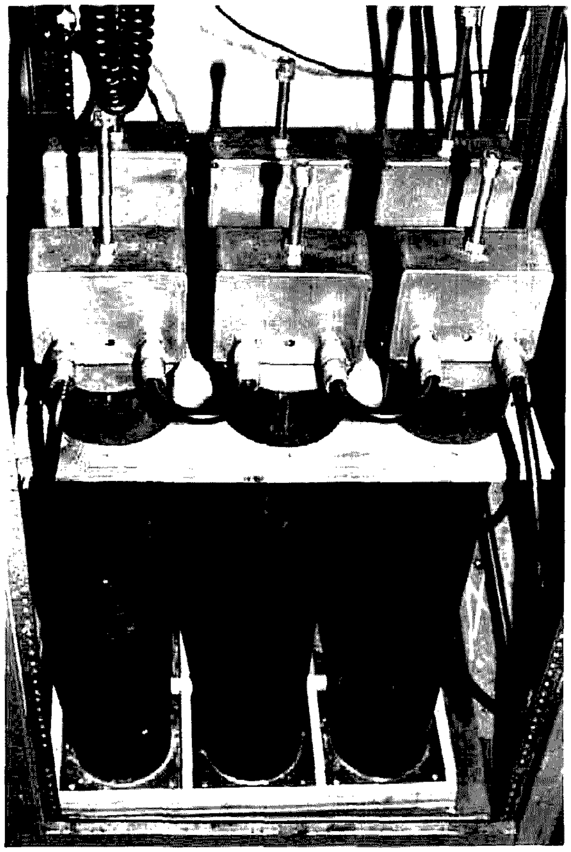


fig. 1. Six-cavity duplexer with a measured notch of 95 dB and an insertion loss of 1.5 dB.

By J.S. Gurske, K9EYY, R.R. 2, Box 178A, Lodi, Wisconsin 53555

After building a repeater, I soon realized that I'd need a duplexer in order to use the same antenna for the repeater, transmitter and receiver. After pondering the situation, I decided to build a six-cavity duplexer similar to the one described in the ARRL *FM Repeater Manual.** This duplexer, while not cheap, still costs much less to build than to buy; the total cost should range from about zero to $250.00, depending on how well your junk box is stocked.

The duplexer I built has a measured 95 dB of isolation and 1.5 dB insertion loss. **Figure 1** shows the completed duplexer in operation on the K9EYY repeater. **Figure 2** shows a completed cavity; **fig. 2A**, a cross-section view (less inductor and capacitor).

construction

The first step in construction is to cut the 4-inch (10-cm) copper pipes to a length of 22.5 inches (56.25 cm). If you are using the thin wall variety of copper pipe, handle it carefully to avoid distortion. Square both ends of all six pieces by using a lathe and a steady rest to support your work.

*Available from Ham Radio's Bookstore, Greenville, New Hampshire 03048 (8.50 postpaid).

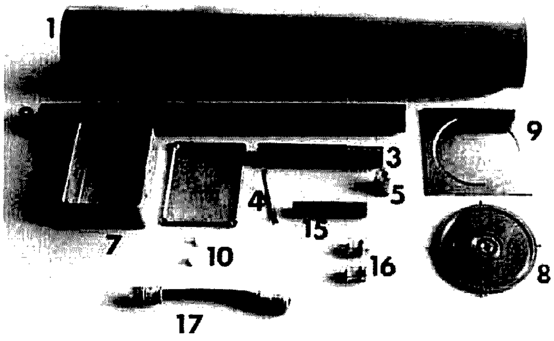
Next cut the 1-3/8 inch (3.44 cm) copper pipes to a length of 18 inches (45 cm). Once again you should square both ends of all six pieces. Cut the 1-inch (2.5-cm) O.D. brass tubing to 6-inch (15-cm) lengths. You will need six of these. Now machine 6 brass plugs from 1-inch (2.5-cm) brass rod stock as shown in **fig. 3**. The center hole in each brass plug is threaded with a 3/8 inch (0.938 cm) × 16 tpi tap.

The teflon insulating bushings are fabricated next. (Refer to **fig. 4**.) Use very sharp cutting tools in the lathe tool holder and rotate the teflon material slowly in the lathe. I used a speed of approximately 35 RPM (I used the back gearing on the lathe) and it cut very easily. You will need two of these bushings for each cavity, for a total of 12 teflon bushings. Be sure to drill a No. 50 hole in each bushing for a No. 16 wire.

Use the lathe to fabricate an aluminum plug (see **fig. 5A**) to be used when silver soldering the threaded rod, the brass plug, and the brass tube. Then fabricate another special aluminum plug to the dimensions shown in **fig. 5B**. This plug will be used to temporarily hold the finger stock inside the 1-3/8 inch (3.5-cm) copper tube while you silver solder the finger stock in place. This plug will prevent the "fingers" from getting too hot and losing their temper. The solder will not adhere to the aluminum.

Thread a nut onto a piece of 3/8 inch (0.938 cm) × 16 threaded rod. Run the nut past the point where you will cut the rod, then cut the rod to a length of 24 inches (60 cm). Then run the nut off the cut end of the threaded rod to chase or clean any threads which may have been damaged when you cut the rod to length. You will need six of these rods (one for each cavity).

Now look at **fig. 6**. Notice that a plastic pipe is slid over the threaded rod and fitted snugly inside the metal box. Check the inside measurement of the metal boxes



Items to be purchased or fabricated for each cavity. The numbers in the photograph correspond to the numbers in table 1.

table 1. Six-cavity duplexer parts list. This is a listing of all the parts you will need to obtain or fabricate to make a six-cavity duplexer. Some of these items are shown in the photograph above.

item number (see photo)	quantity needed	description
1	6	4 × 22-1/2 inch (10 × 56.25 cm) copper tubes
2	6	1-3/8 inch O.D. × 18 inch (3.5 × 45 cm) copper tubes
3	6	1 inch O.D. × 6 inch (2.5 cm × 15.25 cm) brass tubing
4	6	pieces of finger stock to fit inside item 2
5	6	tuning plunger bushing — 1 inch (2.5 cm) diameter brass rod
6	6	tuning rods 3/8 inch (0.95 cm) × 16 threaded rod 24 inches (60 cm) long
7	6	boxes to fit on top of cavities
8	6	top covers for 4 inch (10.5 cm) tubes made of 1/4 inch (0.6 cm) brass plate
9	6	bottom covers for 4 inch (10.5 cm) tubes made of 1/8 inch (0.3 cm) brass plate
10	12	teflon bushings 1/2 × 1/4 inch (1.35 × 0.6 cm)
11	18	nuts 3/8 inch (0.95 cm) × 16 for tuning rods (6 for locking and 12 for tuning)
12	12	coupling loops (made from No. 16 tinned wire)
13	3	inductors (made from No. 16 tinned wire)
14	6	copper straps 1/4 × 1 inch (0.6 × 2.6 cm) No. 0.020 copper
15	6	3/8 inch I.D. × height of mini boxes plastic pipe to keep mini box from compressing
16	12	"N" type chassis coax connectors to couple one cavity to another
17	4	7 inch (17.5 cm) tip to tip RG-192 double-shielded coax to couple the middle cavities to those on each end
18	1	9 inch (22.5 cm) RG-192 double-shielded coax to couple receive cavities to "T" connector
19	1	26 inch (65 cm) RG-192 double-shielded coax to couple transmit cavities to "T" connector
20	3	15 pF small variable capacitor (Johnson 189-5-5)
21	2	lengths of RG-192 to reach from transmitter to transmit cavities and receiver to receive cavities

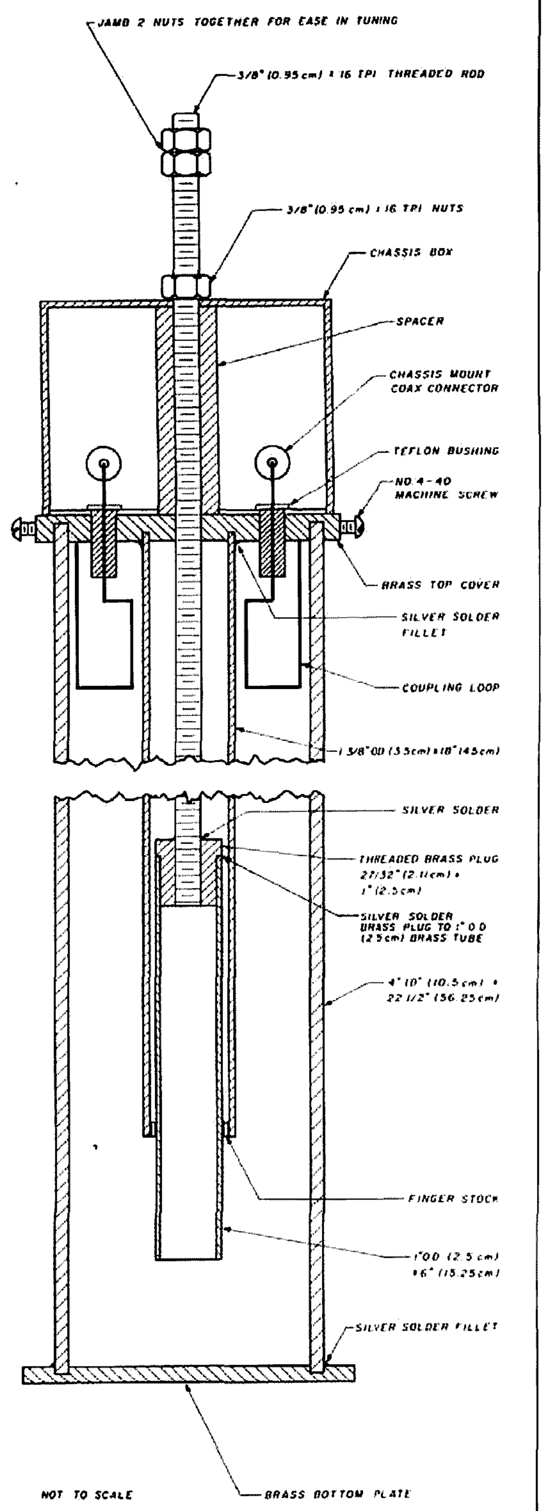


fig. 2A. A cross-sectional view of a single cavity. Neither the inductor nor the capacitor is shown in the metal chassis box.

you will be using and cut six lengths of this 3/8-inch (0.938-cm) I.D. plastic pipe to fit snugly inside the box. This spacer is used to keep the box from changing its shape when you tighten the lock nut after you have adjusted the cavity.

See **fig. 7A**. Make six bottom covers for the six 4-inch (10-cm) tubes by cutting square pieces of 1/8-inch (0.313-cm) brass plate so that they measure 4-1/2 × 4-1/2 inches (11.25 × 11.25 cm). Chuck these pieces of brass in the lathe one at a time and cut a 4-inch (10-cm) slot 1/16 inch (0.175 cm) deep, so that the 4-inch (10-cm) tubing fits snugly into the 1/16 inch (0.175 cm) circle you cut into each 1/8 × 4-1/2 inch (0.313 × 11.25 cm) square piece of brass base. You should have approximately 1/4 inch (0.625 cm) between the circle and the outside edge of these pieces of brass.

Refer to **figs. 7A** and **fig. 7B**. Fabricate six top covers to fit on the 4-inch (10-cm) copper tubes, using 1/4-inch (0.625-cm) brass plate stock. Cut rough 5-inch (12.5-cm) circles from the brass plate stock with a hacksaw. Then insert the rough sawed blank into your lathe chuck. (A 4-jaw chuck might be easier to

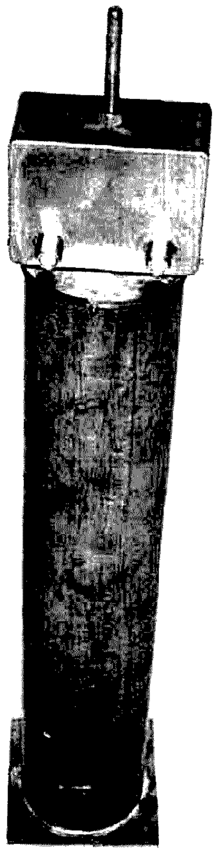


fig. 2. Completed cavity. Note the locking nut on the threaded rod to hold the tuning adjustment. Also note the metal chassis box, the two "N" type chassis connectors, top and bottom covers.

use at this point.) Cut this blank to 4-1/2 inches (11.25 cm) in diameter and drill a 5/16 inch (0.78 cm) hole in the exact center of this cover plate. Thread this hole with a 3/8 inch (0.94 cm) 16 tap. Cut a circular slot 4 inches (10 cm) in diameter and 0.150 inch (0.375 cm) deep. The width of the slot should equal the thickness of the large 4 inch (10 cm) copper pipe, and the slot large enough so that the cover will fit snugly on the top end of the large copper pipe. Cut another round slot 1-3/8 inch (3.5 cm) in diameter and 1/8 inch (0.31 cm) deep to accommodate snugly the 1-3/8 inch (3.5 cm) O.D. copper pipe. Drill 2 holes 3/8 inch (0.94 cm) in diameter exactly 2-5/8 inches (6.56 cm) apart. The centers of these two holes should be exactly 1-5/16 inch (3.28 cm) from the center of the hole you drilled and threaded. These holes will accommodate the teflon bushings.

Refer to **fig. 7B** and notice the four small screws pointing inward toward the center of the cover. These are used to hold the brass covers on the top of the large copper pipes. Use a No. 43 drill and cutting oil to drill 4 holes as close to the bottom of the cover plate as you can. Tap these holes with a 4-40 tap. (Use a good grade of cutting oil or you will break the tap every time. The broken taps cannot be removed and the exposed edges must be ground off.) Fit 4-40 bolts in each of the tapped holes. If you grind a small point on the end of these bolts, they will hold the cover on the pipe more securely.

Refer again to **fig. 6** and also to **fig. 8**. Try to select a cast mini-box rather than a box made by bending sheet aluminum. The cast box will be more rigid and will keep the cavity tuned. Each box will include:

• input and output coax connectors (chassis mount) located 7/8 inch (2.19 cm) up from the bottom of the box, 1-15/16 inch (3.28 cm) from the center of the box.

• a 7/16 inch (1.09 cm) hole through the top and bottom of the box.

• two 3/8 inch (0.940 cm) holes for mounting the teflon bushings; they must align with the holes in the top cover plate.

• two holes for fastening the box to the top plate of each cavity (6-32 bolts).

• a spacer installed between the top and bottom sides of the box to keep the box from distorting when tuning is completed and the lock nut is tightened. This spacer can be metal or plastic. The inside diameter should allow the 3/8 inch (0.94 cm) threaded rod to slide inside the spacer.

The boxes I used measured approximately 4-3/4 inches (11.88 cm) wide, 3-1/2 inches (8.75 cm) high and 2-1/4 inches (5.63 cm) deep and were obtained at a surplus outlet.

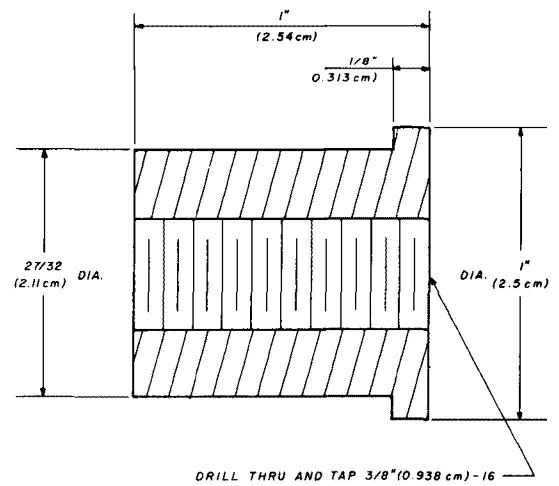


fig. 3. The dimensions of the six threaded brass plugs made from 1 inch (2.5 cm) brass rod.

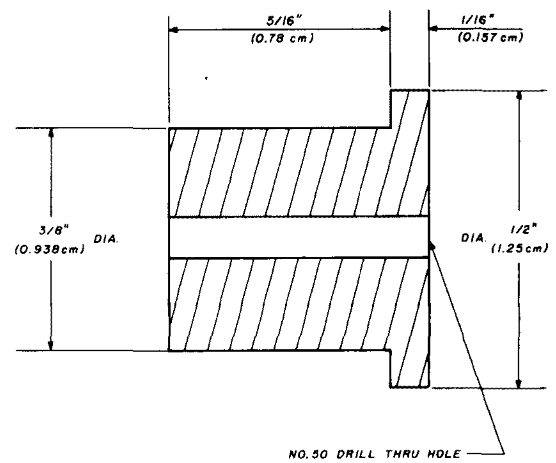


fig. 4. The dimensions of the 12 teflon bushings that allow the coupling loops to pass from the cavities through the top covers and into the chassis boxes. Note the No. 50 drill hole through the center of each bushing.

assembly

Figure 9 shows the main 4-inch (10-cm) diameter copper tube silver-soldered to the brass bottom plate. (Using low-temperature silver solder, we were able to attach the 4 inch (10 cm) tubes to their bases with the heat from only one acetylene torch in spite of the great conductivity of the 4 inch, 10 cm, copper tube.) Place the tube in the slot you machined in each square brass plate, apply flux, and silver solder the base plate to the 4 inch (10 cm) copper tube. Check for trueness before you lay the piece aside.

Next place the 1-3/8 × 18 inch (3.44 × 45 cm) copper tube into the slot previously cut in the round 1/4 inch (0.63 cm) brass top plate. Check for trueness

and silver solder in place. See **fig. 10** which shows the tube silver-soldered to the brass top cover plate. It also shows how the finger stock fits inside the other end of this tube. The finger stock should be silver soldered to the inside of 1-3/8 inch (3.44 cm) tubing at the lower end. **Figure 10** shows its location and how it must contact the 1-inch (2.5-cm) tube for adjustment purposes. When silver soldering the finger stock, do not overheat the fingers. The aluminum plug you made earlier will help prevent overheating. I used a propane torch and the low temperature silver solder mentioned above. Obviously, you will not have the 1 inch (2.5 cm) brass tubing inside the finger stock while silver soldering. Instead, use the aluminum plug to hold the finger stock securely inside the 1-3/8 inch (3.44 cm) copper tube while you are soldering. After soldering, slide the 1-inch (2.5-cm) brass tube inside the finger stock. If the finger stock does not make firm contact with the 1-inch (2.5-cm) tube, remove the

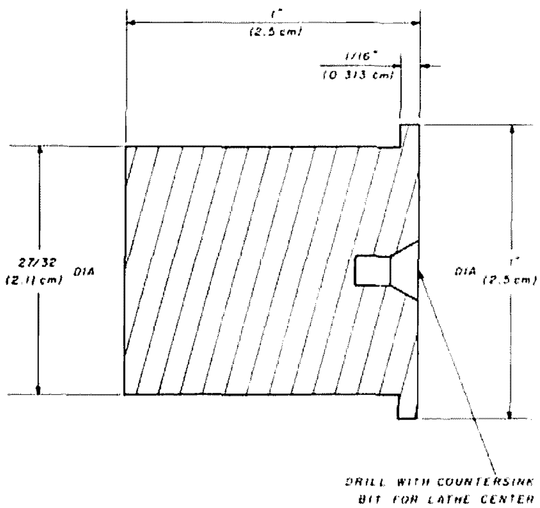


fig. 5A. The dimensions of the single aluminum plug used to fit into the open end of the 1-inch (2.5-cm) tube and supports the open end in the lathe center while the threaded rod, the brass plug detailed in fig. 3, and the brass tube are silver soldered.

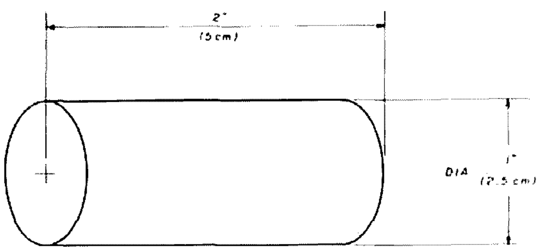


fig. 5B. The dimensions of another aluminum plug to be used when silver soldering the finger stock inside the six 1-3/8-inch (3.44-cm) copper tubes.

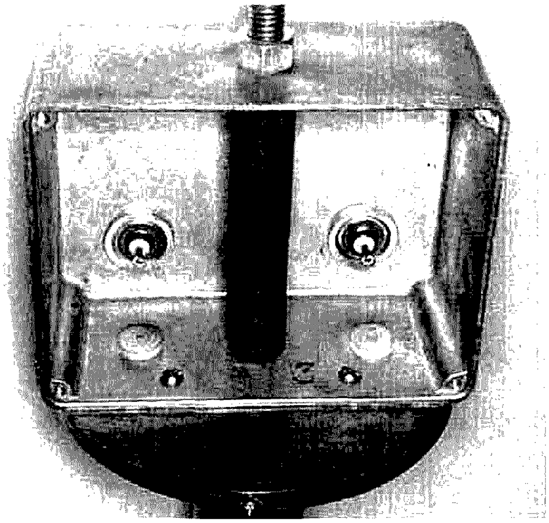


fig. 6. Detail of plastic pipe spacer, threaded rod, locking nut, teflon bushings, coax fittings and how they fit in the box.

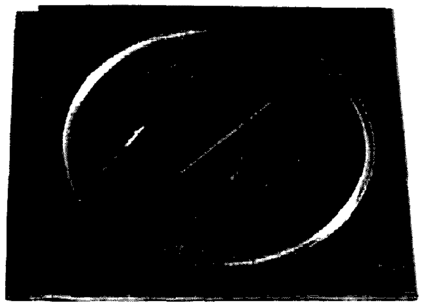


fig. 7A. One of the six bottom covers showing the slot. The large copper pipe will fit in this slot and be silver soldered to the plate.

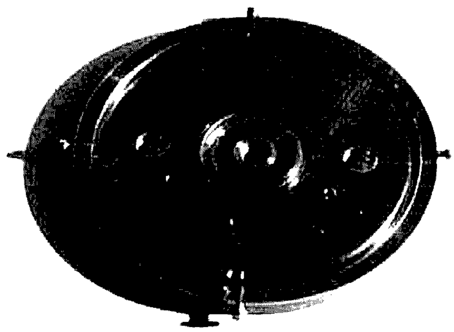


fig. 7B. A completed top cover. Note the center threaded hole, the two holes for the teflon bushings, the two slots for the copper tubes, the 6-32 threaded holes and the four 4-40 bolts in their holes.

1-inch (2.5-cm) tube and bend each finger inward so that it does make firm contact.

Examine **fig. 11** for the following details. It is very important that the 1-inch (2.5-cm) brass tube be attached to the 24-inch (60-cm) length of 3/8 inch (0.938 cm) by 16 threaded rod as accurately as possible. In other words, when the 3/8 inch (0.938 cm) by 16 (adjustment) threaded rod is turned, the 1-inch (2.5-cm) brass tubing should not wobble in the finger stock inside the 1-3/8 inch (3.44 cm) copper tubing. An easy way to accomplish this is to put one end of the threaded rod in the lathe chuck and thread the other end into the brass plug you machined earlier. Slide the 1-inch (2.5-cm) brass tube onto this plug. Then put the small aluminum plug (see **fig. 5A**) in the open (other) end of the 1-inch (2.5-cm) brass tube. Support this end by having the live center (or dead center) ride in the countersunk hole drilled in the center of this plug. The other end of the brass tube is held "centered" by supporting it in the steady rest. Lubricate the steady rest jaws and rotate the entire assembly while silver soldering the threaded rod into the threaded brass plug, and the plug to the brass tubing. When the silver soldering is complete, continue to let the piece rotate until it has cooled. **Figure 11** shows the threaded brass plug, the threaded rod, the aluminum plug, and the 1-inch (2.5-cm) brass tube as well as the steady rest as it was set up in my case.

After the assembly cools, thread the rod up through

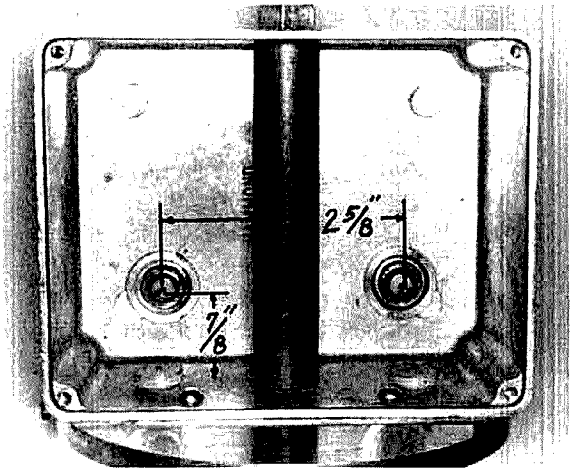


fig. 8. One of the six chassis boxes. These boxes are held to the top cover plates with the two 6-32 bolts in the foreground. Also note the spacer inside the box, the "N" type coax chassis mounts and the teflon bushings.

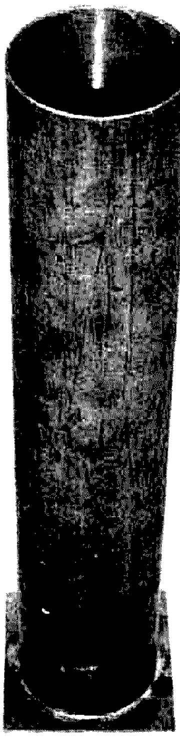


fig. 9. One of the six main copper tubes silver soldered to the bottom cover plate.

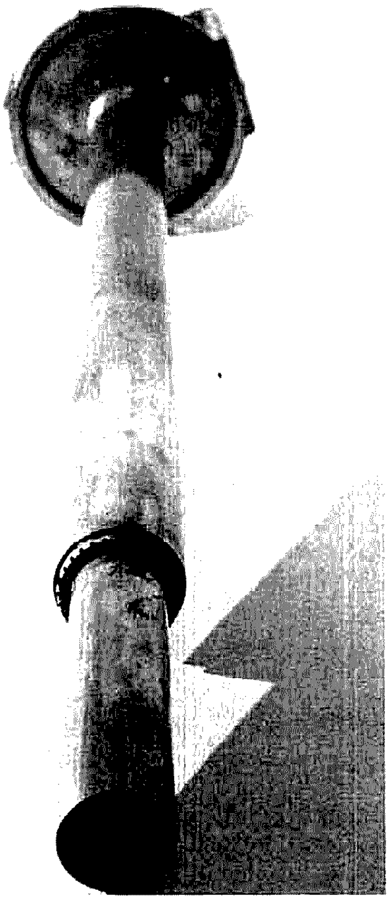


fig. 10. The top plate silver soldered to one end of the 1-3/8 inch (3.44 cm) O.D. copper tube and the finger stock silver soldered to the other end (foreground). Note how the finger stock firmly contacts the inner tube.

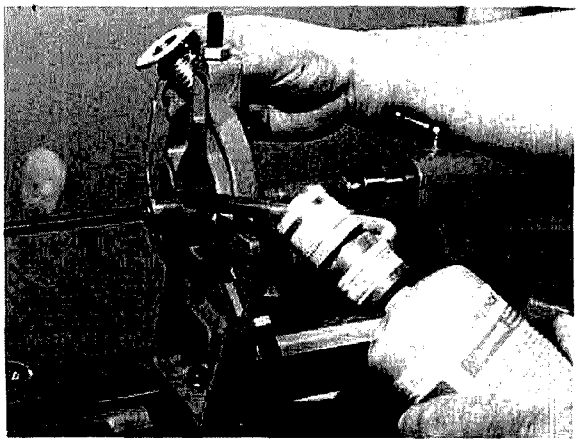


fig. 11. This is how the threaded brass plug is silver soldered to the brass tube and threaded rod while turning in the lathe. Note how the aluminum plug described in *fig. 5A* is used to support the other end of the brass tube in the tail stock live center rest.

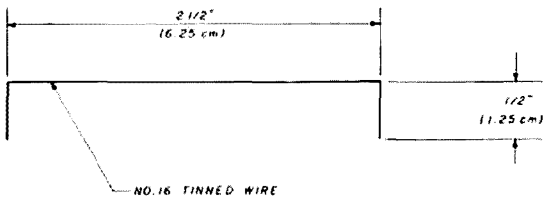


fig. 12. The dimensions and shape of the inductor, which spans the input and output coax chassis connectors inside three of the chassis boxes.

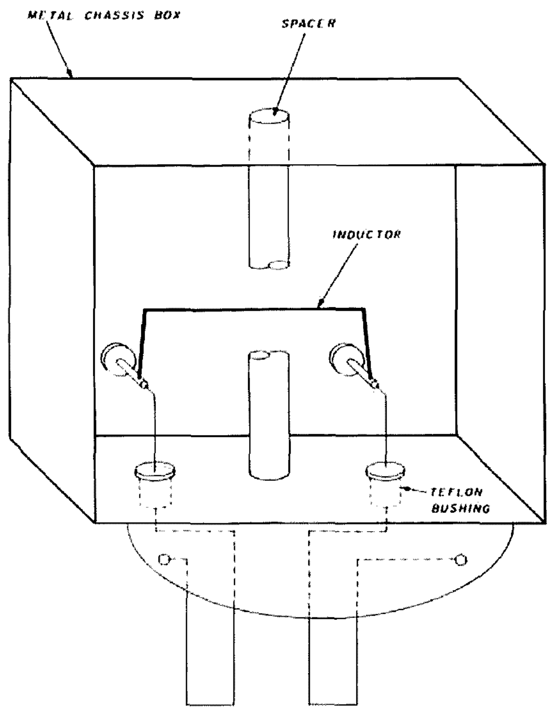


fig. 13. The inductor is connected in three of the cavities.

the brass top covers. Make sure the finger stock firmly contacts the 1-inch (2.5-cm) brass tube. (See **fig. 10**.)

coupling energy to the duplexer

Three cavities use short lengths of No. 16 wire while the other three use a small capacitor. The three cavities that have inductors made of No. 16 wire connected from the input to the output coax chassis connectors will go to the receiver and be tuned to provide a notch at the transmitter frequency.

Refer to **figs. 12** and **13**. Bend a length of No. 16 tinned wire to the dimensions and shape shown in **fig. 12**. Temporarily connect all three inductors between the input and output coax chassis connectors. *Do not solder these wires at this time because you will have to remove them for the first step in the tune-up procedure later.*

The other three cavities will each have a small variable capacitor (15 pF) connected between them. These three can then be put in the transmitter line and the notch tuned to the receiver frequency. The capacitors are connected to the input and output coax connectors with copper strips measuring 1/4 × **1** inch (0.625 × 2.5 cm). Refer to **fig. 14** and cut six pieces of copper flashing. Then connect them to the 15 pF

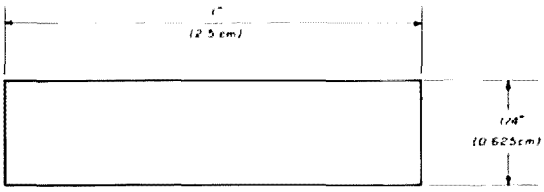


fig. 14. The dimensions of one of the two copper straps that attach to each 15 pF capacitor. Make *six* of these.

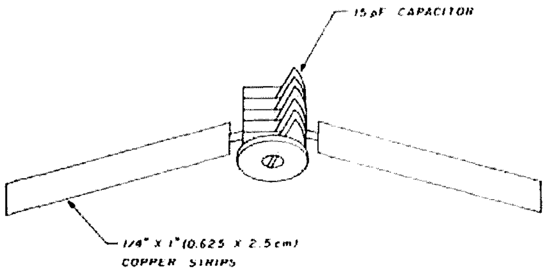


fig. 15. 15 pF capacitors are connected with two copper straps in each of three cavities.

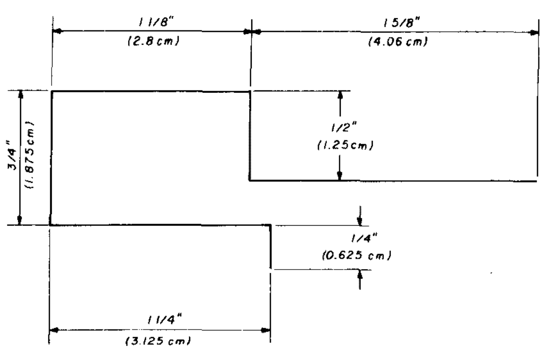


fig. 16. Two of these loops are usd to couple the signal into and out of each cavity. Make 12 out of No. 16 tinned wire. (These measurements are critical, as is the shape of the loops.)

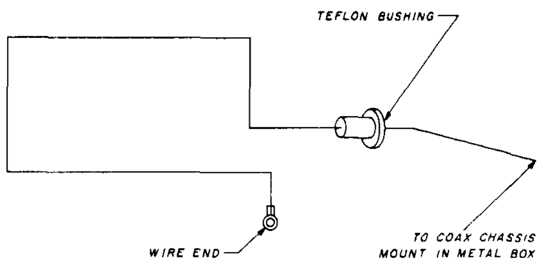


fig. 17. This diagram shows how the wire end is fastened to one end of the loop and how a teflon bushing fits on the wire.

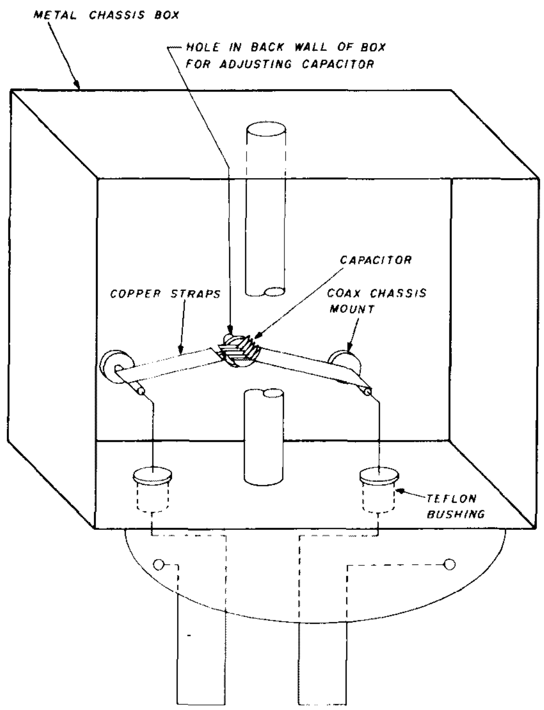


fig. 18. Location of capacitor, copper straps, and coax connectors. Note hole in back wall of box for tuning the capacitor.

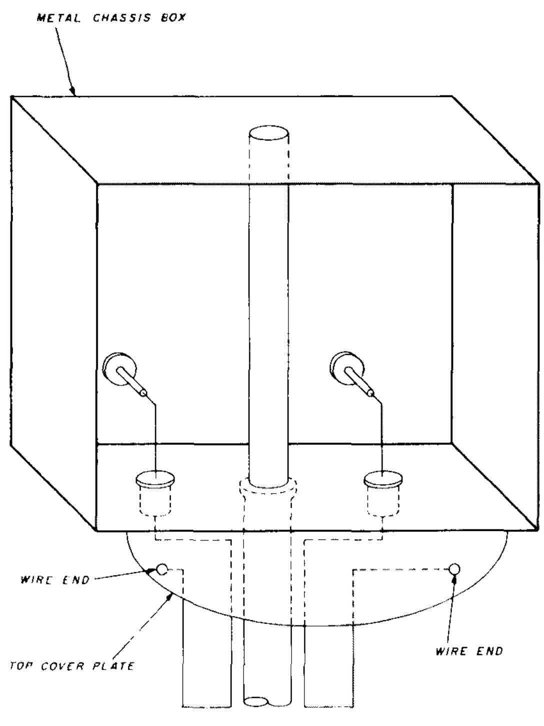


fig. 19. The loops fit on the top covers, pass through the teflon bushings and on to the coax chassis fittings.

capacitors as shown in **fig. 15** and **fig. 18**. Solder these three capacitors to the copper strips (see **figs. 15** and **18**), but do not solder to the input and output connectors yet.

The loops that couple the energy into each cavity are also made of No. 16 tinned wire. *Their shape and dimensions are critical*: use **figs. 16, 17, 18,** and **19** for the proper configuration and measurements. Connect these loops to each input and output connector and to the bottom side of the top plates. Do not solder the wires yet. Hold the copper strips in place and solder the wire coupling loops and simply tack solder the copper strips to the input and output connectors (**fig. 18**).

One end of these loops is tied to the underside of the top plates. I modified some small wire ends and tapped 4-40 holes. The 4-40 screws through the wire ends provide a mechanically secure and electrically good anchor to the underside of the top plates. (See **fig. 16, 17, 18,** and **19**.) The coax chassis mounts are located 5-5/8 inches (6.56 cm) apart, (or 1-5/16 inches, 3.28 cm, on each side of center) 7/8 inches (2.19 cm) up from the bottom of the mini-box.

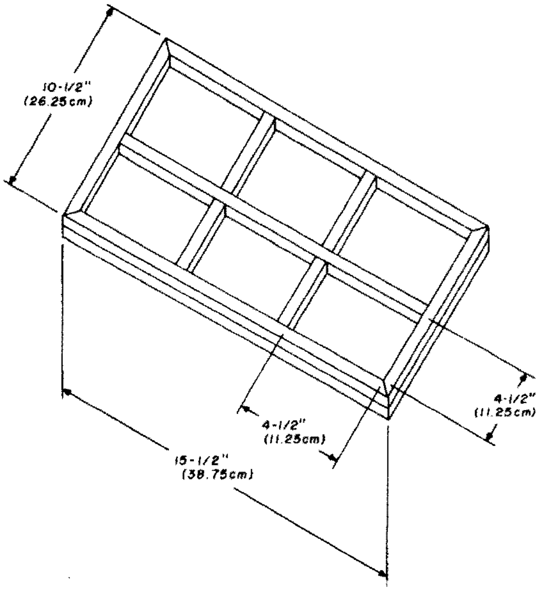


fig. 20. Wooden base holds all six cavities. The base covers fit into the square pockets.

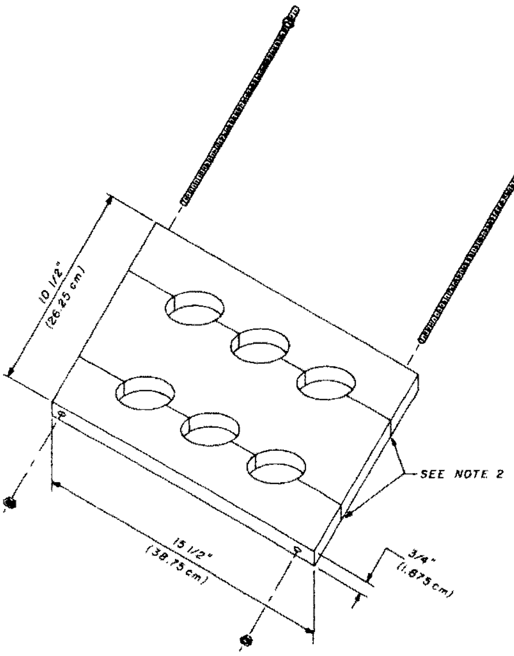


fig. 21. The details of the upper clamp.

building the wooden holder

After all six cavities have been built, mount them together in a holder and tune them to the desired frequencies. (Refer to **figs. 20, 21,** and **22** for construction of the wooden holder clamp assembly.)

The cavities should not be allowed to touch each other as this will tend to detune them. You can build a special holder to prevent them from contacting each other, by following these steps:

Cut a piece of plywood 1/2 inch (1.25 cm) or 3/4 inch (1.875 cm) 10-1/2 × 15-1/2 inches (26.25 × 38.75 cm). Cut 1/2 × 1/2 inch (1.25 × 1.25 cm) wood strips and screw and glue them to the plywood base so that six squares measuring 4-1/2 × 4-1/2 inches (11.25 × 11.25 cm) are formed. This is the base.

Cut a piece of 3/4 inch (1.875 cm) plywood to 10-1/2 × 15-1/2 inches (26.25 × 38.75 cm). Place six marks on one side of this plywood. The six marks should align with the exact center of the six square compartments in the base piece. Using these six marks as the centers of six circles, use a compass to draw 4-inch (10-cm) circles around each of these six points. (See **fig. 21.**)

Mark two straight lines through both groups of three circles. Saw along these lines. You will have three pieces of plywood, each with one-half of three 4-inch (10-cm) circles drawn on the pieces. Using a bandsaw, cut out the 4-inch (10-cm) circle halves. Then drill a hole through the ends of the three pieces of plywood to accommodate a 1/4-inch (0.625-cm) threaded rod. This becomes the upper support. (See **fig. 21.**)

Place a cavity in each base compartment and put at least two wood screws through the brass plate base and into the wood base. Place the upper support around the six cavities and tighten the nuts on the threaded rod.

alignment

We will align the cavities in two stages: stage one for rough tuning each of the six cavities, and stage two for fine tuning the cavities and connecting them all together.

Note that the 3/8-inch (0.94-cm) × 16 threaded rod is for **PASSBAND** tuning. The capacitor, in the case of the transmitting cavities, adjusts the **NOTCH**. (In the case of the receiving cavities, the inductor adjusts the notch.)

In these examples, 147.825 MHz will be used as the repeater transmit frequency and 147.225 MHz as

the repeater receive frequency. We will adjust the three transmit cavities to pass 146.825 MHz and notch the 147.225 MHz frequencies. The three receive cavities will be tuned to pass 147.225 MHz and notch 147.825 MHz.

stage one: rough tuning

Remove the inductors and capacitors from all cavities. Remove all interconnecting coax cables. (See **fig. 23.**) Connect an RF signal generator (amplitude-modulated for convenience) to one of the cavities (**fig. 26**). If your signal generator output is 50 ohms, you will *not* need to use a 3-dB pad. If the output is not 50 ohms or if it is suspect, then connect a 3-dB pad between the signal generator and the cavity. See **fig. 24** if you do not have a 3 dB pad.

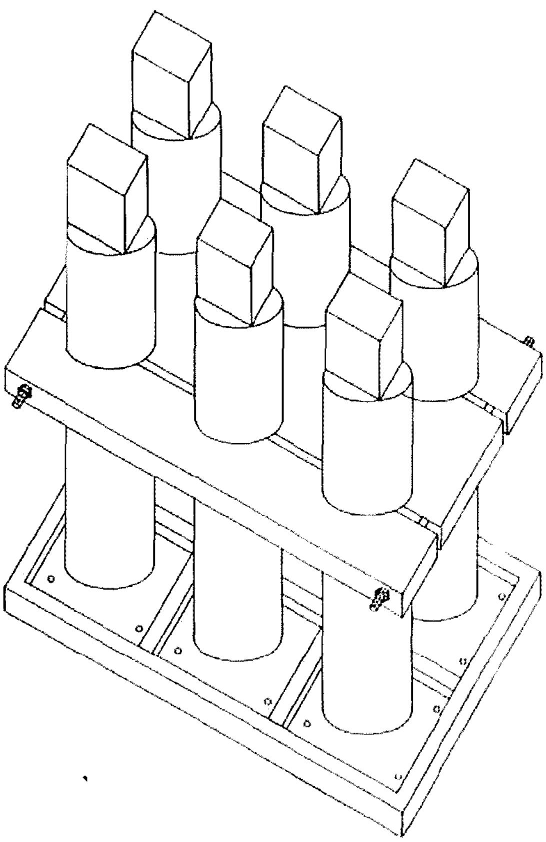


fig. 22. Overview of the wood cavity supporting device.

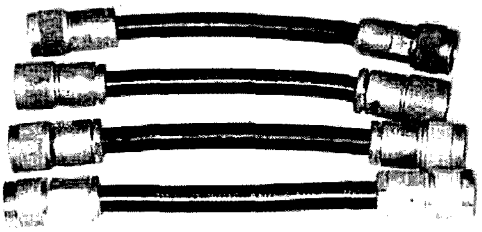


fig. 23. Four cables connect the two middle cavities to the outside ones. Made of double shielded coax, they measure exactly 7 inches (17.5 cm) from tip to tip.

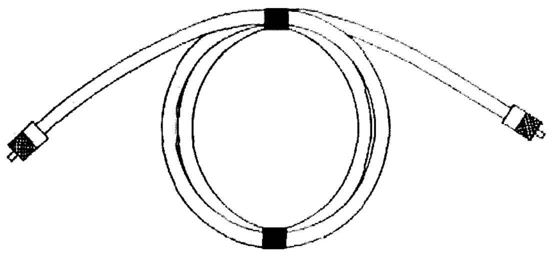


fig. 24. If you do not have access to a 3-dB pad, you can use this device which will be suitable for tuning the cavities. Use RG-58 coax and coil up a 50 to 75 foot (15 to 22.5 meters) length into a neat coil. Put connectors on each end. Tape the coil neatly.

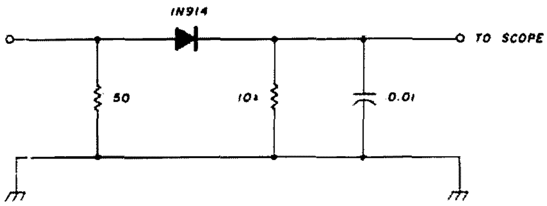


fig. 25. Schematic diagram of a 50-ohm detector suitable for use in this tuneup application. These components should be put in an RF-tight box or enclosure.

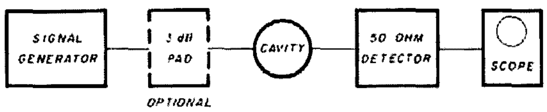


fig. 26. These are the connections for the initial tuning of the six cavities. If you don't have a 3-dB pad, see fig. 24.

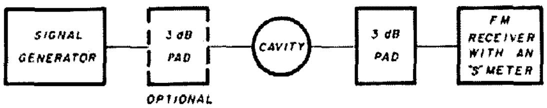


fig. 27. These are the connections to use when setting the notches in the three transmit cavities by adjusting the 15-pF capacitors, and the inductor when adjusting the three receive cavities.

Next connect a 50-ohm detector between the other cavity terminal and an oscilloscope. If you do not have a 50-ohm detector, you can construct one as shown in **fig. 25**.

For the three transmit cavities (the ones which will have the capacitors connected later), adjust the signal generator to within ±100 kHz of 147.825 MHz (**fig. 26**). Then adjust the center threaded rod (passband) until the scope shows maximum energy transfer. You will need to reduce the level of the signal generator as well as the scope gain as tuning progresses.

Adjust all three cavities. From this point on, *leave the threaded rods alone.* Connect the variable capacitors across the input and output connectors of all three transmit cavities.

Connect the signal generator through a 3-dB pad to one of the transmit cavities — now roughly adjusted — then through another 3-dB pad to an FM transceiver which has an "S" meter as shown in **fig. 27**. *(If you do not have a 3-dB pad, refer to* **fig. 24**.) Adjust the signal generator (CW mode) to exactly 147.225 MHz and an S6 reading on the receiver, which is also tuned to 147.225 MHz. Adjust the capacitor (which you just connected in the cavity) for the lowest S-meter reading possible. You may have to increase the output of the signal generator to maintain a visible S-meter indication. When you have obtained the lowest S-meter reading — i.e., the deepest notch — go on to the next cavity.

After all three transmit cavities have been adjusted,

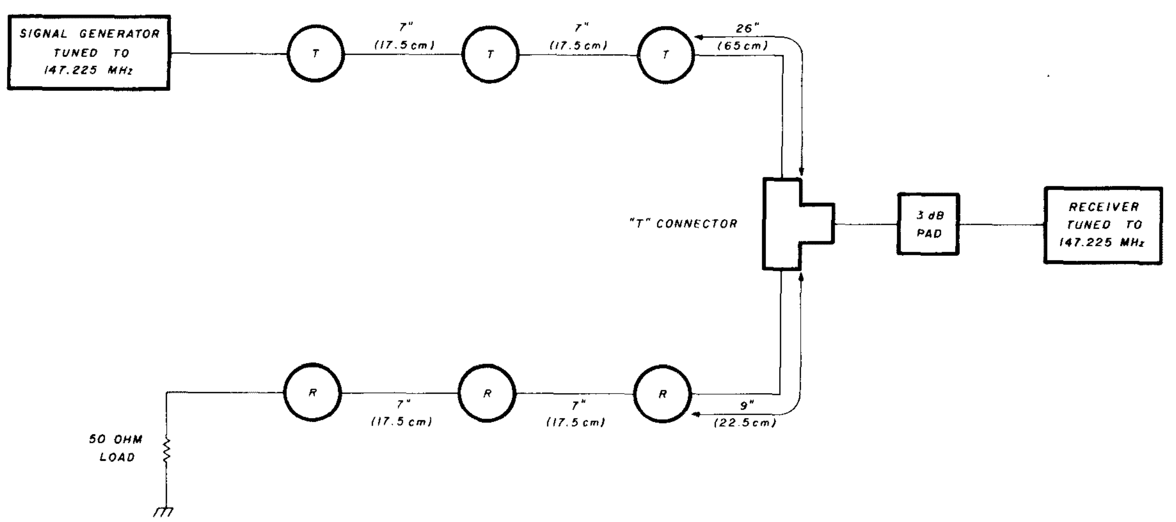


fig. 28. These connections are made for the final adjustment of the duplexer transmit cavities.

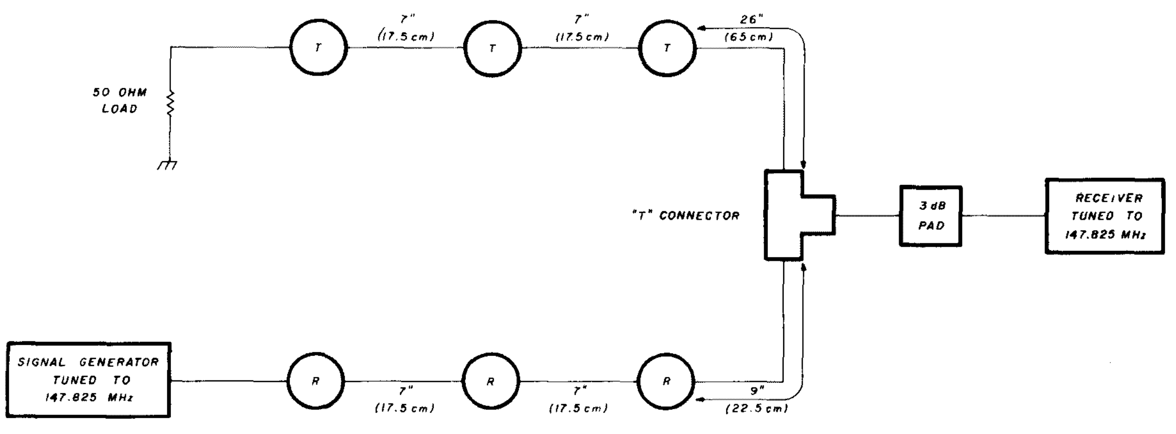


fig. 29. These connections are made for the final adjustments of the duplexer receive cavities.

go back to the beginning of this section and perform the same steps for the three *receive* cavities. Note that you will now be dealing with the inductors you made from the No. 16 tinned wire instead of the capacitors. Adjust the No. 16 wire inductors to deepen the notch. Remember that the frequency to be passed is now 147.225 MHz and the notch frequency is 147.825 MHz.

The inductor dimensions given should be all right, but if the notch is not good enough, try using larger or smaller wire, or even a copper strap if necessary.

stage two: fine tuning

Connect cavities according to **fig. 28.** Tweek the center-threaded rods on the *transmit* cavities only for minimum signal at the receiver (maximum notch). These adjustments interact somewhat. Keep increasing the signal generator output as the S-meter reading decreases. Reconnect according to **fig. 29.** Tweek the center threaded rods on the RECEIVE cavities only for minimum signal (maximum notch). Repeat these steps several times. You may be amazed (as I was) to see the notch get deeper and deeper with each repetition of these steps. You will also notice that bringing objects into the near vicinity tends to detune the cavities slightly as you approach the – 100 dB point.

Lock the threaded rods by tightening a 3/8 inch (0.94 cm) × 16 nut against the mini-box as you finish the last adjustment on each cavity.

acknowledgements

Naturally when one becomes involved in a project of this magnitude, friends often prove helpful. In this regard, I would especially like to thank Ted Gisske, K9IMM, and Chuck Forster, WA9ACI, for their technical help over many months, and Mel Seamans, WB9PKH, for helping with coax when my budget was really strained. I would also like to thank Jim Osborn and Sherm Fusch for their photographic efforts and advice. Joe Androfski, K9OMF, provided constant encouragement, and hands-on help over many hours during the construction and tune-up phases.

I will try to answer readers' questions; please enclose an SASE with any correspondence.

obtaining the parts

Because some of the materials needed are rather unusual, they may not be available at your local hardware store. Others will have to be machined.

The two most unusual items are the pieces of 4-inch (10-cm) diameter **copper drain pipe** and the **brass plate.** The copper pipe can often be found at a large plumbing wholesale supply house, often buried under some other pipe. Try looking under "Brass" in the business section of any large city telephone book; you may find the names of outlets for brass plate listed there. If the ones you call don't stock it, ask for the names of other companies. I found my brass plate through just such a referral. I had to drive about 200 miles, but I got it from Howard Brass and Copper in Milwaukee. (Ask for "tag ends" — they're cheaper.)

Finger stock can be constructed, but the commercial material is much better. I got mine (Stock No. 134B) from Tech-Etch Inc., 45 Aldrin Road, Plymouth, Massachusetts 02360.

The best choice of **solder** for this project is silver solder with a low melting point — i.e., 400 degrees (204 degrees centigrade). Ordinary silver solder, with its higher melting point, is more difficult to work with, and regular hard solder will cause problems if you should decide to silverplate your project later (see "Safe, Sensible Silverplating," page 29). Silver plate will adhere to silver solder, but is likely to flake off of hard solder. My 1/16-inch (0.16-cm) diameter silver solder (manufactured by the J.W. Harris Co., Inc., 10930 Deerfield Road, Cincinnati, Ohio 45242) came from a local rock shop. Do purchase paste or **liquid flux** to use with your silver solder.

The 3/8-inch (0.94-cm) × 16 threaded steel rod can be obtained at almost any large hardware store.

The **teflon** can be found in some hardware stores or plastic stores. I got mine from a friend in the paper business. When large piles of paper stock are cut, a guillotine type of cutter is used. The blade comes to rest against a square piece of telfon. From time to time the teflon is rotated to expose a new unused surface. After the four sides have been used, the strip is replaced with a new one. The used strip has plenty of stock left to make the feed-through bushings.

The **No. 16 tinned wire** is a standard item available at hardware stores and ham swapfests. The small copper strap can be made from a piece of **copper flashing.** Get a scrap from a roofer or builder. All 12 "N" type **chassis mounts and connectors** were purchased at flea markets or swapfests. **RG-192 coax** can be found at dealers or hamfests, as can **Tee connectors.**

ham radio

safe, sensible silverplating

Stake your claim to recycled silver lost in photoprocessing

Have you ever completed a ham radio project and wished to improve its appearance or performance with silverplating? When I finished the duplexer described in the previous article, I wanted to silverplate it. But my previous experiences with electroplating were not encouraging. I knew I'd have to work with silver cyanide, a highly poisonous solution that emits cyanide gas, which can cause illness and even death. I would also have to obtain an expensive silver rod to use as an anode.

To devise a safe, economical alternative, I turned to black-and-white photography, which employs vast amounts of silver in the manufacture of film and printing papers. Even though much of the silver freed in processing is more often discarded with the spent solutions than recycled, it can be reclaimed. I reasoned that if it were possible to recover the silver from spent solutions, namely fixer, or "hypo" — then it should also be possible to capture that silver on a copper tube.

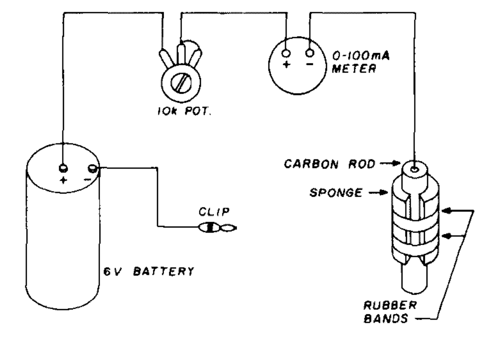


fig. 1. Basic connections for silverplating with photographic fixer.

getting started

The first step in silverplating with reclaimed silver is obtaining an ample supply of exhausted fixer. Sources include photography labs or stores, graphic arts firms that make blueprints or photographic enlargements and reductions, printers with graphic arts departments, and the photography departments of schools and colleges. Friends who process their own film and print their own black-and-white pictures are also good sources.

The best fixer (for your purposes) is that which has been well used; used fixer carries a greater amount of silver than fixer used only slightly. Consequently, your best source of spent fixer may be the least fastidious photographer.

Once you've acquired your solution, obtain a simple dry cell. (I used a 7-year old 6-volt lantern battery.) You'll also need a 10k potentiometer, a 100 mA movement meter, and a carbon rod. (My carbon rod, salvaged from a discarded No. 6 dry cell, measured about 0.75 inches — 20 mm — by about 5 inches — 137.5 mm.)

Wrap the carbon rod in an ordinary kitchen sponge. Secure the sponge with rubber bands. You are now ready to silverplate.

silverplating your project

1. Connect the battery, potentiometer, carbon rod, and meter as shown in **fig. 1**.

2. Carefully clean the items to be plated. You will probably want to use fine steel wool and trisodium phosphate. Both of these materials are available in the paint

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section of large hardware stores or in a paint store.

Once the item is clean, *do not touch it with your fingers*. The oils on your skin can contaminate the surface, ruining an otherwise effective cleaning job.

3. Connect the minus clip to the item to be plated and the plus clip to the sponge-wrapped carbon rod. (See **fig. 2.**)

4. Pour some fixer into a glass or plastic bowl. *Do not use a metal container.*

5. Dip the sponge-wrapped carbon rod into the bowl of *used fixer and rub the* sponge-covered rod along the surface of the item to be plated as shown in **fig. 3**. At the same time adjust the pot for a reading of from 50 to 100 mA. (50 mA is a good choice.) As you rub the sponge-covered rod along the surface of the item to be plated, you will see the silver begin to collect on the item you are plating.

6. With a little practice, you'll soon be able to evaluate the uniformity and thickness of the silver plate. When you are satisfied, go on to the next piece to be plated.

7. Should you want to plate a large, long tube, simply put the sponge on a long stick and slide it inside the tube. (This is simpler and safer than filling the tube with cyanide solution and then inserting a long silver anode into the tube [**fig. 4**]).

At this point I want to caution you about "flash" plating, which can occur if you don't actively guard against it. "Flash" plating occurs when metals that are

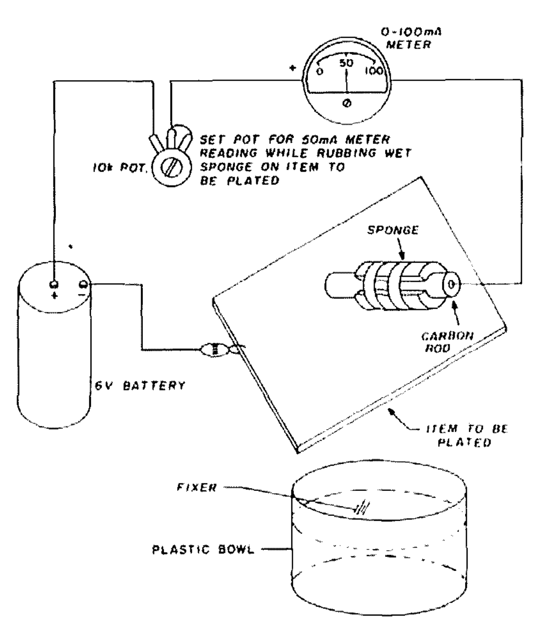

fig. 2. Silver plating with photographic fixer. Note basic connections and how the fixer-soaked, sponge-covered carbon rod is rubbed along the surface of the item to be plated.

fig. 3. Silver plating the tuning rod of a cavity.

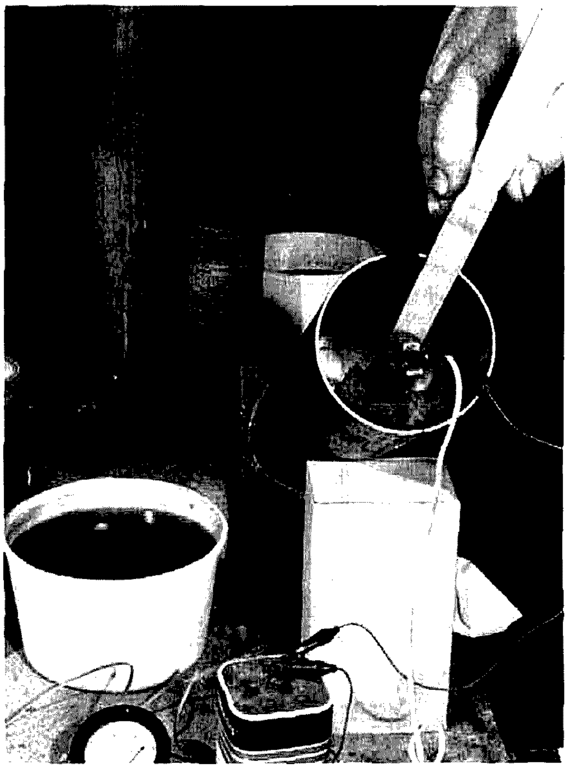

fig. 4. Silver plating the inside of the main tubing of a cavity.

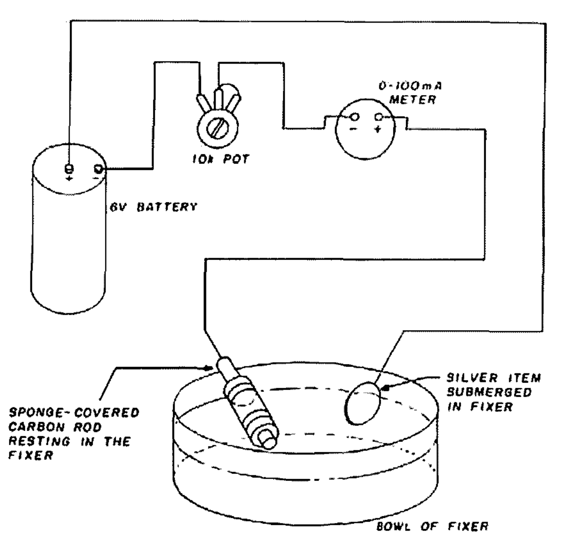

fig. 5. Rejuvenating the depleted fixer: set the potentiometer for a meter reading of approximately 75 mA and wait for about 15 minutes.

high on the activity scale (silver and copper, or silver and brass, for example) come in contact with each other *in the absence of a flow of current.* "Flash" plating looks good but is not permanent. You can prevent it by making sure that the battery is connected every time the item to be plated comes in contact with the fixer-soaked sponge. This is the purpose of connecting the battery (as shown in **step 3** above) *before* you poured the fixer into the bowl.

If your fixer begins to weaken before you've completed the job, but you don't have any more fixer on hand, try this method of rejuvenating the solution. Find an unwanted silver item such as an old vase or piece of discarded flatware. (I used an old silver coin which had been in a fire and was all bent out of shape.) Connect the unwanted silver item to be plated and *reverse* the battery connections. (The meter will read backward unless you reverse its lead. See **fig. 5.**) Place both the silver item and the carbon rod in the fixer solution. Then adjust the potentiometer for a 75 mA reading on the meter. The silver will turn black. After about 10 or 15 minutes, you can resume silverplating for a while longer but *don't forget to reverse the battery and mA meter leads.* I don't know how long or how many times you can go through this rejuvenation cycle. I did it about six times and it seemed to work fine.

I'm sure you'll obtain good results if you follow these suggestions. My cavities looked 100 percent better after they were plated and I know they work better, too. So, go ahead and silverplate your next *ham radio* project.

ham radio

the
weekender

programmable call sign identifier

This programmable identifier — originally conceived by Don Henry, W3FE — can be used in a number of applications including home station or repeater ID and remote link identification. When Don suggested that a 64-bit shift register could be used to store a call sign, I proceeded to develop a circuit that would accomplish this task, be easy to program, and contain few IC's.

construction

The circuit contains four CMOS IC chips (**fig. 1**). CMOS technology is well suited to this application because of its extremely low quiescent current requirements, thus making an on/off switch unnecessary. Because the circuit is powered all the time, backup power for the programmed data is unnecessary.

Circuit layout is not critical; whatever is convenient should work. I used a 4-1/2 inch (11.5 cm) by 4-1/2 inch (11.5 cm) piece of Vector Board with 0.1 inch (2.5 mm) hole spacing for holding the components. The resistors and capacitors associated with U4 (4093) should be kept as physically close to the IC as possible. I recommend sockets for holding the ICs. Wire wrappng works well; point-to-point wiring could also be used to make the connections. The power supply requirements are simple; a single 9 volt alkaline Duracell (or equivalent) battery will do.

programming

The programmable identifier has two modes of operation controlled by the DPDT switch. One mode is *Recirculate* or playback and the other is *Program*. The programming sequence is as follows:

- Set the DPDT mode switch to the Program position.
- Push the reset switch.

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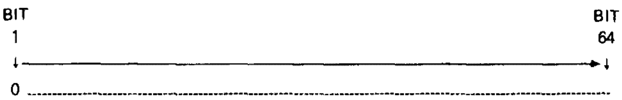
table 1. Programming data chart.

character	bit pattern	number of bits
A	1 0 1 1 1	5
B	1 1 1 0 1 0 1 0 1	9
C	1 1 1 0 1 0 1 1 1 0 1	11
D	1 1 1 0 1 0 1	7
E	1	1
F	1 0 1 0 1 1 1 0 1	9
G	1 1 1 0 1 1 1 0 1	9
H	1 0 1 0 1 0 1	7
I	1 0 1	3
J	1 0 1 1 1 0 1 1 1 0 1 1 1	13
K	1 1 1 0 1 0 1 1 1	9
L	1 0 1 1 1 0 1 0 1	9
M	1 1 1 0 1 1 1	7
N	1 1 1 0 1	5
O	1 1 1 0 1 1 1 0 1 1 1	11
P	1 0 1 1 1 0 1 1 1 0 1	11
Q	1 1 1 0 1 1 1 0 1 0 1 1 1	13
R	1 0 1 1 1 0 1	7
S	1 0 1 0 1	5
T	1 1 1	3
U	1 0 1 0 1 1 1	7
V	1 0 1 0 1 0 1 1 1	9
W	1 0 1 1 1 0 1 1 1	9
X	1 1 1 0 1 0 1 0 1 1 1	11
Y	1 1 1 0 1 0 1 1 1 0 1 1 1	13
Z	1 1 1 0 1 1 1 0 1 0 1	11
0	1 1 1 0 1 1 1 0 1 1 1 0 1 1 1 0 1 1 1	19
1	1 0 1 1 1 0 1 1 1 0 1 1 1 0 1 1 1	17
2	1 0 1 0 1 1 1 0 1 1 1 0 1 1 1	15
3	1 0 1 0 1 0 1 1 1 0 1 1 1	13
4	1 0 1 0 1 0 1 0 1 1 1	11
5	1 0 1 0 1 0 1 0 1	9
6	1 1 1 0 1 0 1 0 1 0 1	11
7	1 1 1 0 1 1 1 0 1 0 1 0 1	13
8	1 1 1 0 1 1 1 0 1 1 1 0 1 0 1	15
9	1 1 1 0 1 1 1 0 1 1 1 0 1 1 1 0 1	17

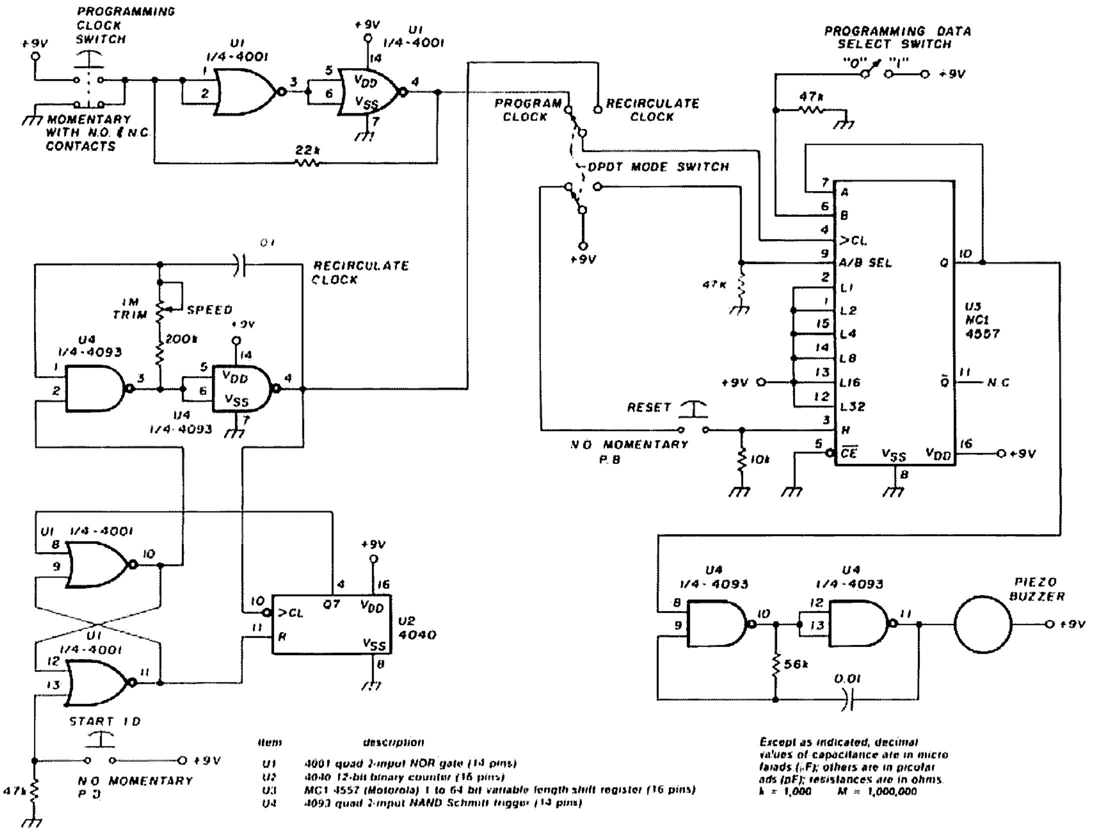
Note: The first bit is always a 0. Insert 0 0 0 (3 bits) between characters.



- Set the programming data select switch to the "0" position.
- Push the programming clock switch once — Bit 1 is not programmed (note: Bit 1 must always be a "0".)
- Set the programming data select switch to the appropriate "0" or "1" position.
- Push the programming clock switch once.
- Repeat previous two steps until all 64 bit positions are programmed.

After the 64th bit is programmed, set the mode switch to the Recirculate (or playback) position. Push the start ID switch and the programmed data will be played back.

More Details? CHECK—OFF Page 128



item	description
U1	4001 quad 2-input NOR gate (14 pins)
U2	4040 12-bit binary counter (16 pins)
U3	MC1 4557 (Motorola) 1 to 64 bit variable length shift register (16 pins)
U4	4093 quad 2-input NAND Schmitt trigger (14 pins)

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

fig. 1. Programmable identifier uses only four ICs.

To program your call, follow the Programming Data Charts (**table 1**) to get the correct data bit sequence for each letter and number. Be sure to program 0 0 0 (3 bits) between each character to allow spacing between them., Keep count of the number of bits programmed. If your call does not require all 64 bits, insert zeros for the remaining bits up to and including the 64th bit.

If you want to use the programmable IDer with other equipment, a few points need to be mentioned. To start the IDer a positive-going pulse on U1 pin 13 of the 4001 will accomplish this. The Morse output can be taken from the *Q* output of U3 pin 10, the MC1 4557. The 1 megohm trimpot is used to adjust the recirculate (playback) speed.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

high power amplifiers: part 2

In last month's column, I discussed the design of high power VHF/UHF tube amplifiers with emphasis on the effects of the new FCC Amateur power regulations.[1] This month's column will continue in the same vein, but focus instead on the construction and practical application of VHF/UHF amplifier designs.

DC circuitry

In designing and building high power VHF/UHF power amplifiers, most Amateurs concentrate their efforts only on the RF portion of the design. This is unfortunate because power supply voltages and currents — as well as DC circuitry — are often just as important for successful amplifier RF performance.

Sometimes Amateurs will pay hundreds of dollars for a high power transmitting tube and then neglect to review the manufacturer's specification sheet before firing up the new bottle in the amplifier. Instead, they'll ask "Joe Blow" what voltages he uses on the same tube and follow his advice. Ironically, these Amateurs wouldn't purchase used gear if the instruction manual weren't part of the deal.

Suppliers of high power tubes — such as Amperex, Varian/EIMAC, General Electric, and RCA — all provide comprehensive data sheets with their products. Usually 4 to 12 pages long, these data sheets list mechanical dimensions and RF/DC characteristics. They also specify recommended bias and operating parameters so that the user can obtain the maximum performance and lifetime possible from the tube. Some manufacturers also offer, at a nominal price, abbreviated specification books listing many different tube types.[2,3] The ARRL *Handbook* also includes tube tables.[4]

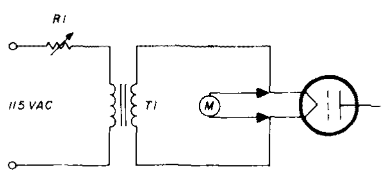


fig. 1. Recommended filament voltage circuit for high power transmitters. The voltage at the tube pins is set by placing an adjustable wirewound resistor, R1, in series with the primary winding. If the dropping resistor is not used, the transformer, T1, must have a current limit close to that required by the tube. Note that the voltmeter, M, is connected directly to the tube socket pins. See text for other precautions and recommendations.

Some of the more significant low-frequency or DC parameters for high power tubes are the maximum, minimum and recommended operating voltages and currents, warm-up time, and grid (and screen grid if a tetrode is used) supply impedance and plate dissipation.

filaments

Filament voltage in the newer tubes is, unlike the old receiving type tubes, far from standard. Because of the high power involved, the filament current required is often quite high. As a result, tubes may have a significantly shorter lifetime if the filament voltage is not kept within the manufacturer's recommended ±5 percent values.

Let's take a few examples. The popular 4CX250B has a filament rating of 6.0 ±0.3 (±5 percent) volts at 2.6 amperes. *Filament voltage must be measured at the tube socket pins with an accurate (within 1 to 2 percent) AC voltmeter with the tube inserted and drawing filament current!* 6.3 volts AC is a common transformer (voltage) used on receiving tubes. Such transformers are often used on transmitting tubes (for example the 4CX250B), but they usually have some means of lowering the voltage. This is most commonly done with an adjustable wire-wound resistor placed in series with the primary of the filament transformer.

Now, let's examine the higher power 8877, which requires a filament voltage of 5.0 ±0.25 volts (±5 percent) at nominally 10 amperes. Since 5.0 volt transformers are quite common, they are usually used *but rarely with any series resistor in the primary since it looks like one isn't needed!*

When a high power tube is first turned on, the filament resistance is always quite low. Initially, a high input current surge occurs; this can significantly shorten tube life. In the case of the 4CX250B, the resistor (which is typically used in the transformer primary) will serve to limit the turn-on current surge. However, in the case of the 8877, if no such primary resistor is used, the surge current will increase in a manner limited only by the resistance of the transformer secondary winding. Furthermore, no adjustments (in primary resistance) would be possible if the line voltage were already lower than nominal.

Therefore, when selecting a filament transformer it's best to use one that is slightly higher in voltage than required so that an adjustable power resistor can be placed in series with the primary winding (**fig. 1**). Another alternative is to place a Variac™ on the primary side of the filament transformer and then increase the voltage slowly on each initial turn-on while monitoring the voltage directly at the tube socket. Many commercial trans-

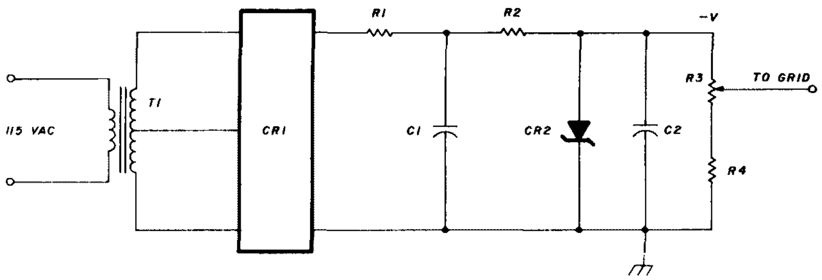


fig. 2. Recommended control grid supply for high-power amplifiers. T1 is chosen to supply the required voltage. CR1 may be a full wave or bridge rectifier. R1, C1, and C2 are filter components. CR2 is a zener diode or VR tube whose current is set by R2. R3 is a 1000-ohm wirewound potentiometer. R4 can be eliminated if the regulated voltage is low (less than 50 volts) or 500-1000 ohms if the voltage is greater than 50 volts. See text for further description.

mitters have a built-in filament voltage regulators to compensate for power line variations.

If you do not use a primary limiting resistor, be certain to always use a filament transformer with a current capability close to but not much higher than the filament current rating of the transmitting tube, selected to enhance current limiting. This may explain why some Amateurs when using filament transformers with higher current ratings than required, have experienced premature filament burnout of high-power transmitting tubes.

Each tube has a minimum warm-up time before any cathode current should be drawn. This can vary from as short as 30 seconds to as much as five minutes for the indirectly heated high-power transmitting tubes! Some manufacturers specify a lower standby filament voltage (for example, −10 percent) that can be instantly brought up to specification during transmit times, thereby extending tube life. This can be easily accomplished with an inexpensive relay activated by the station send/receive control line. More information on such operation can be found on the manufacturer's comprehensive data sheets, so consult them before you begin operation. Use of a time delay relay is also recommended in the high voltage supply.

Another VHF/UHF phenomenon, especially in grounded grid amplifiers or those where the cathode is hot, is RF on the filament of the tube. This can cause RF current to flow between the cathode and the filament, resulting in increased input drive requirements, additional filament heating, and possibly reduced tube life. To prevent this, the filaments should be bypassed to the cathode and connected to the filament transformer through a bifilar choke.

At VHF and especially UHF, a problem called transit time heating often occurs because of the finite time required for the electrons to move from the cathode through the grid to the plate, where some electrons may be repelled. This problem can usually be controlled by decreasing filament voltage at higher operating frequencies. Recommended filament voltages at various frequencies are provided on the manufacturer's data sheets. Additional information on this subject is contained in reference 5.

grid bias

In the past, triodes were usually biased by grounding the cathode and applying a negative bias voltage to the grid. The new high-μ triodes are usually easier to bias. Often the grid is directly grounded and a large high-power zener diode, rated at the required bias voltage, is placed in series with the cathode circuit. However, many VHF/UHFers use only simple grid supplies often consisting of just a rectifier diode and filter capacitor on a reverse connected filament transformer. The problem with this approach is that if there is any grid current, the grid voltage will fluctuate directly as a function of the impedance of the bias supply.

Good design practice requires the grid supply to be regulated. In addition, the voltage should be adjustable and have a maximum DC output impedance between 1 and 2 kilohms! This can be done with either a shunt type transistor (or tube) regulator or a wirewound potentiometer directly across the output of a regulated grid supply, as shown in **fig. 2**. The main advantages of a low impedance grid bias supply are decreased grid drive and improved IMD.

screen supply

The majority of tetrode amplifier configurations require a screen voltage supply. Over the years I've seen dozens of complex schemes with separate regulated supplies, current protection relays, and surge voltage protectors. Most of these techniques are unnecessary. Screen grid burnout can occur if the screen voltage is not decreased rapidly when plate voltage is removed. Furthermore, negative screen current can occur in a properly operated tetrode amplifier. I believe that the most foolproof screen supply voltage circuit is a shunt regulator consisting of a dropping resistor and appropriate voltage regulators shunted to ground and supplied from the tube's plate supply voltage. Furthermore, I highly recommend the use of VR (voltage regulator) tubes rather than power zeners. VR tubes are low in cost (especially at flea markets), easy to use, offer long life and don't require heatsinking and insulated washers such as required by power zeners. VR tubes are less likely to be damaged by current surges from arcs caused by "barnacling" (more on this shortly).

The reverse screen grid current problem on tetrodes is often ignored by Amateurs. It is specifically noted on the manufacturer's data sheets and can be easily handled by placing a properly selected resistor from the screen

grid to ground. In most cases 20-25 kilohms (per tube) is optimum. Don't forget to include this current when selecting the dropping resistor value in the regulator circuit just recommended. The final recommended circuit is shown in **fig. 3**. The circuit is simple and should be all that is required to properly supply voltage to the screen grid in a high power amplifier. In addition, if the series dropping resistor value is properly selected, the tube's screen dissipation can be limited to the specified value.

plate supply voltage

Finally, we must supply the tube plate(s) from a high voltage supply. (This is always the user's choice.) First a transformer must be selected. When running over 500 watts, I highly recommend use of 230 volt primaries since there will be improved regulation without blinking the house lights off and on!

The choice of the rectifier circuitry, either a full wave, full wave bridge, or full wave voltage doubler (all these are acceptable), is a personal choice based on the transformer chosen and its voltage ratings. A common practice is to use a set of high voltage solid-state rectifiers, often feeding just a bank of high voltage capacitors. One should carefully choose the amount of capacitance required before completing the design or use the alternative choke input filter.[6]

Regardless of the configuration chosen, there are several other considerations. The primary of the high-voltage transformer should be adequately fused. An adequate bleeder should be placed across the high voltage output to not only discharge the supply when turned off but to *also* improve regulation. This is especially important with a choke input filter. Recently there has been a tendency to place a high power, high value resistor in the cathode of transmitting tubes when in the standby mode. *I do not recommend this practice but instead prefer to entirely disconnect the high voltage from the final when not in the transmit mode.* This not only prevents inadvertent shocks but also eliminates any noise or self oscillations that may be possibly generated by the final or feedthrough from the earlier stages. This is easily accomplished by using a vacuum relay in the high voltage output lines as shown in **fig. 4**. These relays are often available at flea markets or surplus stores at reasonable cost.

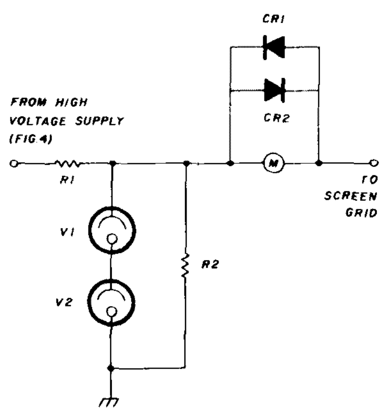


fig. 3. Recommended screen grid supply. The quantity and voltage of the VR tubes, V1 etc. is determined by the tube specifications. R1 should be a high power (200 watts typical) wirewound resistor whose value is selected to provide adequate current to the VR tubes (30 mA typical) as well as R2. R2 should be typically 20-25,000 ohms per tube (see text). CR1 and CR2 are meter protection diodes.

Sometimes there is an advantage to connecting the screen grid directly to ground to stabilize high power UHF amplifiers. This technique requires a slightly different power supply topology as shown in **fig. 5**. If this configuration is chosen, the screen supply must also be able to handle the full current required by the plate. However, the total voltage required from the plate high: voltage supply is reduced since it is in series with the screen supply.

Sometimes a small impurity becomes lodged between the grid or screen grid and plate of a high-power tube, causing a short circuit. This phenomenon is called "barnacling." A limiting resistor should be placed in the high-voltage line so that the peak current does not exceed 50 to 100 amperes and destroy the tube. This is accomplished by placing a 25 to 50-ohm resistor in the plate voltage line. The power rating of the resistor is determined by the plate current of the tube in use. If a barnacling condition occurs, this series resistor will limit the plate current and possibly disintegrate, but the tube will survive and will usually return to normal operation after the initial surge. Remember, if a shunt regulator is used on the screen, put this limiting resistor on the supply side as shown in **fig. 4**. In this way, if the resistor blows open, the screen voltage will also be removed.

A high-voltage in-line fuse should also be used. Low-voltage glass or ceramic fuses are not recommended because they are slow acting, can explode, and can arc, possibly causing damage to meters! A short length of a small gauge wire (No. 40 preferred) can be used as a fast acting, low cost high voltage fuse.[7]

amplifier bypassing

It is sound engineering practice to use an extra pair of coaxial relays so that high power amplifiers may be bypassed when not in operation or when the final is warming up (**fig. 6**). The input relay can be a low power type but the output relay must be capable of handling the full output power of the transmitter.

Amplifier bypass relays should never be (hot) switched when RF power is present since this will severely limit relay contact lifetime not to mention generation of spurious pulse noise. Isolation of each bypass relay doesn't have to be as high as with a typical T/R relay, with 30 dB per relay more than adequate. *Furthermore, particularly in grounded grid amplifiers, RF should never be applied to the amplifier input unless the proper voltages are already applied.*

plate dissipation

This parameter is often overlooked. The plate dissipation is approximately

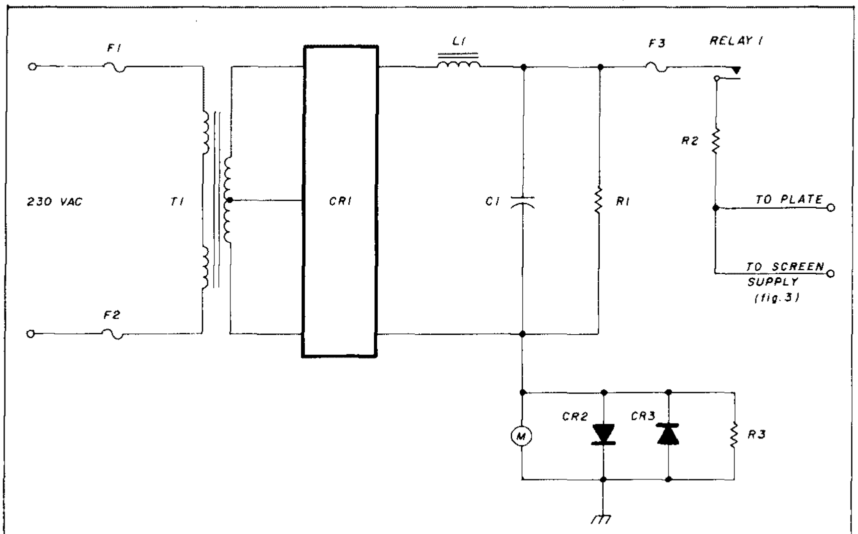


fig. 4. Recommended high voltage supply for a high power amplifier. F1 and F2 are line fuses. F3 is a fine wire described in text. C1, CR1, L1, R1, and T1 are described in text. CR2 and CR3 are meter protection diodes. R2 is described in the text (25-50 ohms typical). R3 is a meter protection resistor (50 ohms 10 watts typical). RL1 is a vacuum relay described in text.

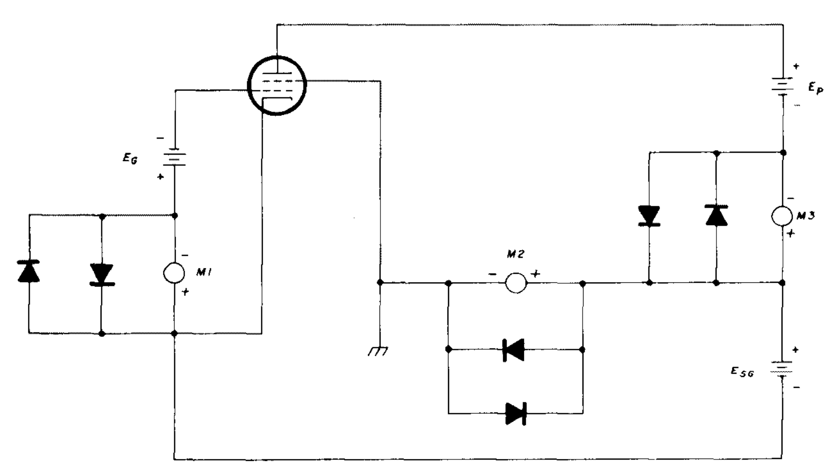


fig. 5. Power supply configuration used with DC grounded screen grids. Power supplies are represented as batteries. M1 is grid current, M2 is screen current, and M3 is plate current. The meters are protected by back-to-back diodes.

the plate power input less the power output of the amplifier (other circuit losses being low). For example, if the plate input power is 2500 watts and the amplifier output is 1500 watts, the plate dissipation is approximately 1000 watts. The plate dissipation ratings of many of the more common tubes is shown in **table 1** of reference 1. (The subject is further discussed in the "cooling" section of this article.)

cooling considerations

Amateurs are notorious for abusing high power transmitting tubes. One of the most misunderstood transmitting tube parameters is the plate dissipation rating. *The manufacturer's specification applies only if there is adequate cooling, by either air or water, to the tube.*

Herein lies a problem. Each air-cooled tube has a different cooling specification, clearly spelled out on the manufacturer's data sheet, based on the *amount of air circulation (in cubic feet per minute) through the plate fins or around internal anode tubes as well as a specified back pressure (usually measured in inches of water).* Therefore an exchange of tubes based primarily on a higher plate dissipation rating, may often require a greater air flow rate — for example, as in the case of swapping the 4CX250B and 8930 in the K2RIW 70-cm amplifier.[8]

External anode tubes are best cooled using air system sockets so that air is circulated over the grid and cathode seals as well as through the plate fins. Since these tubes generally have a high resistance to air flow, especially when compared to internal anode tubes, chimneys are a must to force the air through, rather than around, the plate fins. A large blower, preferably one operating at 3600 RPM, is usually required to keep plate dissipation within the tube ratings and overcome the high back pressure developed by the tube. (Don't forget to occasionally remove external anode tubes from service and clean out any accumulated dirt or dust that becomes lodged in the heat sink fins.)

Gary Madison, WA2NKL, has developed an approximate formula to compare the back pressure of different blowers as shown in **eq. 1**:

$$P_B = 8 \times 10^{-9} d_W^2 V_W^2 \qquad \textbf{(1)}$$

where P_B is the blocked-off back pressure in inches of water
d_W is the wheel diameter in inches
V_W is the speed of the blower in revolutions per minute

For example, a 2-inch wheel diameter 3600 RPM blower will have a back pressure of approximately 0.415 inches of water while an 1800 RPM blower of

the same size would develop only 0.104 inches in water. This formula *dramatically illustrates how a small increase in wheel diameter or speed will* increase the back pressure.

A recent trend is to use water cooling on the plates of transmitting tubes, especially at UHF. For over 10 years I have been using such a technique on a 4CW800B type water-cooled tetrode. My 2-gallon water supply is housed in a small waste paper container; the pumping pressure is provided by a fish tank pump. Although distilled water is preferred, I find that water collected from a de-humidifier is adequate. Each summer I store the required 10-12 gallons needed for a year of operation. Replacement water-cooled plate radiators are now commercially available for the 2C39/7289 family of tubes.

This technique is very good, but not without its limitations. Often the tube's grid and cathode seals are overloaded. *They also require cooling.* In grounded grid configurations the grid, if attached directly to the chassis, may be sufficiently cooled by conduction, but the cathode is usually overlooked. Therefore, if the plate is water cooled, a reasonable amount of air should also be circulated through the cathode/grid compartment.

Often when the proper amount of air is used, there may be an objectionable blower noise in the shack. Some Amateurs have attempted to lower the noise by locating the blower elsewhere and directing the air through a hose. This method requires special attention, since the hose chosen will have resistance, which will decrease air flow. Therefore, if a blower is remotely located, its output may have to be raised accordingly.

There are two main problems of insufficient air flow. The first one is obvious: tube life is shortened. The other may also be obvious but is seldom mentioned: if plate dissipation is too high, the extra heat is often dissipated in the tank circuit, especially when metal tubing type plate lines are used. The net result is that the tank circuit tuning drifts especially during long transmissions such as on EME.

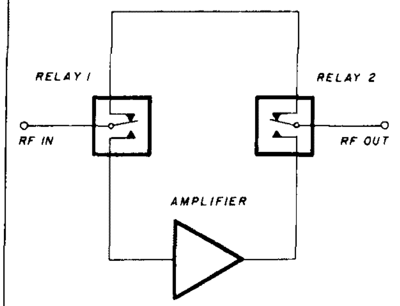


fig. 6. Relay bypassing scheme for protecting a high-power amplifier during warmup. See text for relay description.

Always try to provide a good safety margin (2:1 is recommended) of cooling, whether with air or water. Also verify that there is sufficient pressure differential across the tube, especially for external anode tubes. Also, in the case of air or water supply failure, provide a blower or water flow cutoff switch in the power supply (more on this later). A convenient method of measuring flow and back pressure is described in reference 9.

construction techniques

Many good construction tips are buried in the various articles on high power finals and some of these references will be cited later in this article. Do read as many of the references as possible.

One of the most important design considerations centers on the circuit chosen.[1] The next consideration is the choice of tube type. Choosing a larger tube than required may be wise because the actual plate dissipation will be a smaller percentage of the tubes ratings. However, all is for naught if the mechanical integrity of the amplifier is not given proper attention. All components and mechanical parts should be sturdy and properly secured to the chassis because any changes in temperature or duty cycle will cause detuning affects. Also, the components used in a high performance transmitting amplifier should be first quality. If not, there may be unnecessary down-time, as well as the possibility of destruction to the other components in the amplifier such as the tube! It is not a good practice to *economize* here.

In a high performance, high power amplifier, it is essential that all RF components be placed within shielded boxes (more on this later). It is also wise to locate all other components *outside* the RF enclosures. This is particularly true of the plate compartment. The only items that should be present in the output circuitry are the tube, tank circuit components, RF choke and the high voltage bypass capacitor. Likewise, the input compartment should not contain components such as meters or filament transformers.

The method employed to physically tune the plate circuit is very important for stable, trouble-free operation. No matter what you do there will always be some heat dissipated in the tank circuit. Therefore, it is wise to locate the tuning capacitor where thermal detuning effects are minimal.

A typical example of how to and how not to tune a final is shown in **fig. 7**. Note that in **fig. 7A**, any expansion or contraction of the tank due to heating or cooling causes the capacitance to increase or decrease and therefore detunes the tank circuit. However, if the tuning method of **fig. 7B** is used, the tank circuit will only move back and forth past the tuning capacitor and cause a much smaller change in amplifier tuning.

The tuning capacitors must be carefully designed. A disc soldered on the end of a brass screw is a very lossy tuning method which may cause erratic tuning. I prefer a push mechanism in which there are no moving metal-to-metal contacts. For instance, a beryllium copper flapper can make an excellent tuning capacitor by pushing against it with a non-metallic object such as a threaded teflon rod. W2GN has suggested the use of a metallic rod if an insulator such as a teflon button is placed at the contact point to the capacitor (see **fig. 8**).

Another item to note is that when a transmitting tube has a plate contact ring, the tank circuit should preferably be attached to it rather than to the out-

side of the plate cooling fins. The circulating RF currents in a typical Amateur high power transmitting amplifier can easily reach 10 to 50 amperes!

Finally, there is often a need to provide air inputs and exits. These can best be made using a waveguide beyond cutoff configuration. It is well known that if the diameter of a hole is small with respect to the frequency (much less than 1/4 wavelength) and the hole has depth (such as tubing), RF will be attenuated approximately 30-33 dB per unit length. For example, if a 1 inch (2.54 cm) diameter tube is 1 inch long, it will attenuate VHF signals approximately 30 to 33 dB. If the same tube is extended to 3 inches (7.62 cm), the attenuation will be approximately 90 to 99 dB, usually more than sufficient. This attenuation principle is applied to the air inlet by using either a long large diameter tube, or several smaller diameter shorter tubes placed in a circle for air inlets and exits. Recently a honeycomb type of material has become available and it is easier to use than bundling up tubing, a technique now in wide use.

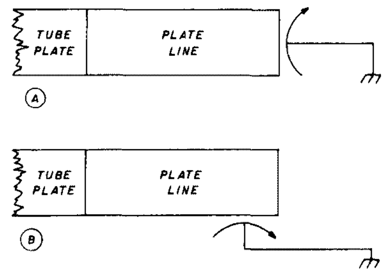


fig. 7. Two methods to tune a plate tank are shown. The method in fig. 7A is not recommended for reasons described in the text.

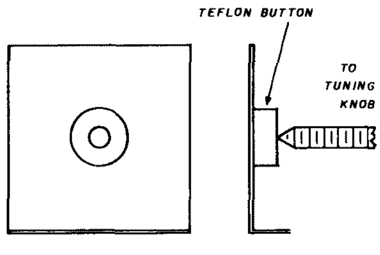


fig. 8. A recommended tuning method for a flapper type of plate tuning capacitor without metal-to-metal contact (see text).

tube selection

So far I have covered many of the different aspects of high power RF amplifier design but haven't said too much about the actual tube selection. **Table 1** in reference 1 shows some of the more popular tube types. Now that the FCC power regulations have been changed to apply to output power there are really many choices.

However, high power transmitting tubes should never be operated above their maximum frequency of operation *unless the voltages and currents are reduced accordingly.* Furthermore, the efficiency usually drops past the rated frequency and hence there may be severe over-dissipation not only in the plate but also in the control grid elements. Unfortunately, at VHF and UHF frequencies, control grid dissipation may increase but will not necessarily show up on the grid current meter as it does at HF. *Failure to observe the tube's ratings can severely decrease efficiency and the life of the tube!*

As I mentioned before, the use of parallel tubes should be a last resort. Using tubes above their rated frequency is also not recommended. If you require only moderate output power (500 watts or less), there are lots of single-tube amplifier choices such as the 4CX250B, 4CX300A, 8930, 8874, or the new 3CX800A7.

There is an alternative to the parallel tube approach. If additional power is required, identical amplifiers can be summed using hybrid power splitters and combiners. However, this technique is tricky and is recommended only for the more experienced VHF/UHFers.

The 4CX350A is not recommended above 135 cm (220 MHz) since it is primarily an HF tube and will exhibit low efficiency at 70 cm. Furthermore, its operating voltages must be lowered above 30 MHz. If you look carefully at the specification sheet you will see that the plate radiator is similar to the 4CX250B but the required air and back pressure have been increased to obtain the 350 watts of dissipation, a technique I mentioned earlier in this article.

For the full legal power limit, the new high-μ triodes such as the 8877 and 8938 are great but costly. Despite rumors to the contrary, the 8938 is not the 8877 with a different base! The 8938 cathode lead inductance has been lowered and the transit time loading has been improved, extending full power efficient operation up to approximately 500 MHz. The 8877 should not be used at full ratings above the 135 centimeter band.

The 4CX1000 series of tubes is often available on the surplus market at affordable prices. However, if stable operation on 2 meters or above is planned, the 4CX1000K is recommended since it has improved input-to-output isolation. The 7213/7214 and 7650/7651 tubes, also often found on the surplus market, will work well through about 1000 MHz, but, despite the data sheet, are not recommended for the 23-cm band since gain and efficiency will be so poor (33 percent if you are lucky!) that the useful output power will be low at maximum ratings. High power on 23 cm and up is still a problem, and no moderately priced single tube that can generate over 500 watts of output power is yet available to the Amateur service.

component selection

The type of RF coupling capacitors chosen is very important. The so-called TV doorknob types are not recommended because they cannot handle the typical RF current in a high-power VHF/UHF amplifier. The Centralab 850 or 857 series or the ITT 50 or 58 series transmitting doorknob type of capacitor (or equivalent) is strongly recommended even though they cost more (available from Radio Kit, Box 411H, Greenville, New Hampshire 03048).

HV feedthrough capacitors are also required. The Murata/Erie model

2498/001-XUO-102M 1000 pF at 4000 volts, is relatively expensive but highly recommended since it will handle the typical high voltages required as well as act as an RF low-pass filter. Some designers have "rolled their own" bypasses by placing a slab of copper or brass against a chassis wall insulated by a thin (0.005-0.010 inch or 0.13-0.25 mm) dielectric insulator.[10]

High voltage connectors are a must especially for safety. The typical "BNC" or "N" types have identical breakdown voltages which are too low for most amplifiers and are not recommended for carrying high voltage. However, the "MHV" (high voltage BNC) connector is highly recommended. Because its end is insulated, making it difficult to touch the tip — a useful feature should a high voltage line be accidentally touched when energized. This feature has saved my life more than once!

Chassis and enclosures are becoming difficult to obtain. Byers Chassis-Kit (K3IWK) has a large selection of chassis kits and will make up special enclosures to order.

When dielectric materials are used at VHF/UHF frequencies, careful selection is important for low loss. I don't recommend nylon or bakelite because they're lossy materials. Some of the recommended materials are Kapton, Isomica, mylar, and teflon. The first two are often difficult to obtain and may be expensive. Mylar, a low-loss dielectric, especially above 300 MHz, is often overlooked. In plentiful supply, it is often used for taping printed wiring circuitry and has a high breakdown voltage (5000 volts/mil) and a dielectric constant of approximately 3.

Teflon is recommended. It has a dielectric constant of only 2.1 and a voltage breakdown of only 1000 volts/mil. It also is porous. Hence for best breakdown it should be rubbed with a silicon grease (e.g. Dow Corning DC-4) or used in dual sheets.[10]

operating considerations

Under the new FCC power measuring regulations it makes sense to design VHF/UHF power amplifiers for linear service right from the start. A linear amplifier is more versatile and does not necessarily require bias adjustments when changing operating modes, thus simplifying switching circuitry. However, ALC circuitry should also be considered to improve linearity. Also the high voltage supply regulation usually improves since the transmitter current excursion is reduced.

I'll say it again! Before operating any tube, consult the manufacturer's data sheets! Excessive voltages can cause arcing and tube destruction. Excessive currents, especially in the cathode, can significantly reduce tube life. Many Amateurs tune their amplifiers for maximum smoke, then get indignant when the amplifier or tube goes up in smoke! Always pay particular attention to the cooling and backpressure. It's better to have too much air flow than not enough.

Also, determine the minimum required filament warm-up time before any cathode current is drawn. Properly built and operated power amplifiers don't require frequent tuning adjustments.

When a power amplifier is working properly, it should generate maximum output power when you are tuned to resonance (a point just above the dip in the plate current). On tetrode amplifiers the same should be true and the screen current should peak (even though it may be negative at the time) at the same point. It is highly recommended that a dedicated screen grid current meter be provided because the screen current is a very sensitive indicator of tuning and loading.

safety considerations

Whenever you are working on a high-power amplifier, safety should be your first concern. Safety applies in both the DC and RF area. *High voltage can kill.* All amplifiers should be adequately shielded to prevent inadvertent contact with the high voltage supply lines as well as the RF circuitry. High voltage primary interlocks, bleeder resistors in the supply, and high voltage discharge mechanisms are a must. Use of a relay in series with the high volt-

age (as previously recommended) is one additional way to lessen the chance of high voltage shocks, but even a relay's contacts can stick — so don't leave anything to chance. Meters, especially those connected directly in the high voltage line, are particularly dangerous. High voltage may be very close to the glass and a source of arcing. If a meter is placed in the high voltage line, it should be adequately insulated from ground and placed behind a protective panel. Bypass meters with back-to-back diodes as shown in **figs. 3** and **4**, to lessen the chance of meter destruction at high currents.

The loss of cooling air or water to the tube can also present a safety hazard. Besides endangering the tube, loss of cooling increases the chance of fire or explosion should the plate overdissipate. An air or water flow high voltage cutoff switch is recommended in case of failure.

Finally, we must respect RF and microwave radiation. In recent years it has become increasingly obvious that Amateur transmitters, especially those operating at 500 watts and above, have sufficient RF power to be hazardous, especially to the eyes and brain. Power densities of 10 milliwatts/cm^2 were formerly the American National Standards Institute (ANSI) C95.1-1974 limit.[11] This limit has now been superseded. The new ANSI C95-1-1982 limits are 1 milliwatt per square centimeter from 30-300 MHz, following the curve f(MHz)/300 milliwatts per square centimeter from 300-1500 MHz, and 5 milliwatt per square centimeter above 1500 MHz. *This limit can be easily exceeded within 1 meter of a high power VHF/UHF transmitter with the plate compartment shield removed.* Blackwell and White have described a simple probe to measure RF power density.[12] *A power amplifier should never be operated with the shielding removed, especially on 2 meters and above!*

warranty considerations

Most manufacturers will not replace a defective tube, even under warranty, unless certain parameters have been monitored. These include, but are not limited to, filament voltage, hours of operation (as indicated on a built-in elapsed meter), circuitry and operating voltages. These requirements are usually spelled out on a warranty sheet in the original packing container. It's well worth checking into these requirements before placing a tube in service.

recommended designs

Finally we come to the question of what are the most popular and recommened designs. The following table includes many recommended references or designs, listed by frequency band:

band	see reference
6 meters	13-20
2 meters	19-35
135 cm	36-41
70 cm	10,42-48
33 cm	49

23 cm and up is still a difficult area; it will be covered in a forthcoming column, but reference 50 is recommended for a medium power (100-150 watts) 23 cm-amplifier.

This list should cover most of the tubes and circuits presently in wide use. The selection of amplifier and tube is left to the individual.

The AM-6155, a moderately priced ($150) surplus linear amplifier, has recently become available from Fair Radio Sales.[51,52,53] It comes complete with a self-contained power supply and uses a single 8930 tetrode. Properly modified, this amplifier can operate on either 2 meters, 135, or 70 centimeters. With 10 watts of drive, 400 to 600 watts of output have been reported.

summary

Properly operated tubes used in a good circuit with appropriate attention given to component selection will have a long lifetime and provide many hours of satisfactory operation.

I highly recommend that you obtain a copy of reference 5 for your library. This reference is very complete and will provide amplification on many of the items discussed in this article. Let me know if I have left out any major points. They can be covered in a subsequent column.

acknowledgements

This month's column would not have been possible without the efforts of many other individuals. In particular I want to thank Lewis Collins, W1GXT, and Gary Madison, WA2NKL, who helped a great deal by reviewing this month's manuscript and providing many valuable suggestions.

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important VHF/UHF events in February, 1985

February, 8: ***EME perigee***

Note: If you know of any important VHF/UHF events, please let me know well ahead of time so that they can be listed in this column.

ham radio

defining the decibel

Why bother? Because in electronics — as in any science — definitions *do* make a difference.

The unit of the decibel, or dB, is used quite widely in electronics to express everything from amplifier gains to bandwidth ratios (is that 10 log or 20 log for bandwidths?). Often the question, "Is that dB-voltage or dB-power?" is heard. The answer to that question is an unequivocal "yes." The short article that follows reiterates the apparently long-forgotten history of the decibel and discusses both its proper application and a few of its common misapplications as well.

The decibel is, roughly, the smallest change in acoustic power that the ear can detect.[1] It's one tenth of a bel, a unit named for Alexander Graham Bell, whose original research revealed the logarithmic amplitude response of the human ear;[2] not surprisingly, the concept of the bel was originally used in the field of telephony. But the unit was found to be too large for practical application, and the decibel was soon found to be more convenient.

In the original acoustic terms, the decibel was defined as 10 log to the base of 10 of the ratio of two acoustic intensities (powers). (A similar but much less frequently used unit is the *Neper* — from Napier — which is given as 1 log to the base e of a *voltage* ratio.[2] Yes, the multiplier is 1, not 10.) In modern electronics, however, the decibel is *defined* as 10 log of a power ratio in which the two powers ratioed are measured at a particular point in a system — at the output of an amplifier, for example. *This is the only definition. Other descriptions of the decibel, such as 20 log of a voltage ratio, are derivations of this definition, often with some critical information omitted.*

The decibel is really just a type of mathematical shorthand. It is more convenient, for example, to express the power gain of an amplifier as 80 dB than as 100,000,000 watts/watt. One variation to this basic definition has been a generalization to allow the two powers ratioed to be at *different* points in a system that have equal impedances — for example, the power gain of an amplifier in a constant 50-ohm system expressed in dB as 10 log of the ratio of the output power to the input power. Such a generalization, however, is still consistent with the original definition. Consider a 50-ohm amplifier in a 50-ohm system. Its input and output impedances must both be 50 ohms to be consistent with the 50-ohm system. Therefore, the source driving the amplifier will deliver the same power to a 50-ohm termination as it delivers to the amplifier input. Let us choose the 50-ohm input to a power meter as the point of measurement of the original definition above. First apply the source directly to the power meter input and record the source power. Then, remove the source from the measurement node (power meter input), apply it to the amplifier input, and apply the amplifier output to the power meter. Measure the new power at the point of measurement. The amplifier gain in dB is then 10 log of the ratio of the second measurement, the output power, to the first measurement, the power applied to the input. This measurement technique is a direct application of the definition of the decibel.

Alternately, we could, by some means, measure the input and output powers of the amplifier with it attached to the source and 50-ohm load (computed from measured input and output potentials perhaps) and compute the gain in a similar manner as above. There is a subtle difference between this second measurement technique and the first. In the first, a single point of measurement, the input to the power meter, was used to measure the two powers for the ratio; in the second, two different points in the system were observed — the input and output ports of the amplifier.

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Since the system was defined to be a 50-ohm system throughout, both techniques will yield the same results. However, if the impedances in the system are not the same throughout, the results will *not* be the same. (This will be demonstrated later.) So, the ratio of two powers, P_2 and P_1, in a constant impedance system expressed in dB, is given by **eq. 1**, where the term "*D*" is simply a general notation and the form "*dB*" is used to show the units of the result.

$$D = 10 \log (P_2/P_1) \ [dB] \quad (1)$$

In many cases the power, P_1, is chosen as some convenient reference power such as one milliwatt. The value of D is then given in dB referred to a milliwatt, abbreviated dBm. This is still consistent with the basic definition of the dB since D is then a representation of the actual power at a point in a system compared to the chosen reference power at that same point.

Now, we will expand **eq. 1**. However, since this is not a lesson in arithmetic, the impedances will be defined as being real with no imaginary part, which will simplify the math considerably. Each of the powers in **eq. 1** may be expressed in terms of the potentials and corresponding impedances, or resistances for the case of impedances with only a real component, at the power measurement points. Expanding **eq. 1**:

$$D = 10 \log [(E_2^2/R_2)/(E_1^2/R_1)] \quad (2)$$

$$= 10 \log (E_2^2/E_1^2) - 10 \log (R_2/R_1) \quad (3)$$

$$= 20 \log (E_2/E_1) - 10 \log (R_2/R_1) \ [dB] \quad (4)$$

This is an interesting result. We can now see where the 20 log of a voltage ratio expression originated in the widely used dB expressions. But what about the second term in **eq. 4**? Well, in the case of a constant impedance system, the two resistances are the same value, resulting in a second term of 10 log(1), which is of course zero. As a result, D is correctly expressed in dB as 20 log of the ratio of two voltage measurements, which is the expression so familiar to many of us. So we'll make a note of it here:

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$$D = 20 \log (E_2/E_1) \ [dB] \quad (5)$$

This would be a good point to digress a moment and examine the question, "Is it dB-voltage or dB-power?" Consider a 50-ohm amplifier in a constant 50-ohm system. If we applied an input signal of one microwatt (− 30 dBm right?) to this amplifier and measured an output power of one milliwatt (0 dBm?), what would be the gain of that device in dB? Letting that power ratio be represented by "G" and applying the basic definition of the dB given in **eq. 1**:

$$G = 10 \log (P_o/P_{in}) \ [dB]$$

$$G = 10 \log (1 \text{ milliwatt}/1 \text{ microwatt}) \quad (6)$$

$$= 10 \log (1000) = 30 \ dB$$

We could have easily found this result by *subtracting* the − 30 dBm input level from the 0 dBm output — 0 dBm − (− 30 dBm) = + 30 dB. To show this mathematically, consider the following:

$$G \ [dB] = 10 \log (P_o/P_{in})$$

$$= 10 \log (P_o/P_{in}) + 10 \log (1 \ mW/1 \ mW)$$

$$[\text{add } 10 \log (1 \ mW/1 \ mW) \text{ which } = 0]$$

$$= 10 \log (P_o/1 \ mW) - 10 \log (P_{in}/1 \ mW)$$

$$= P_o \ [dBm] - P_{in} \ [dBm]$$

Now we will try it from a voltage point of view. First we must compute what input and output voltages we would measure with the corresponding powers given. We all know that the power dissipated in a resistor with a potential E applied across it is given by:

$$P_R = E^2/R \ [watts]$$

Solving for E in terms of P and R:

$$E = (P \cdot R)^{1/2} \ [Volts, \ RMS]$$

For the one microwatt input (remember, we said we had a 50-ohm system):

$$E \ (1\mu W) = (1 \cdot 10^{-6} \text{ watts} \cdot 50 \text{ ohms})^{1/2} = 7.07 \text{ millivolts RMS}$$

And for the one milliwatt output:

$$E \ (1 mW) = (1 \cdot 10^{-3} \text{ watts} \cdot 50 \text{ millivolts})^{1/2}$$

$$= 223.6 \text{ millivolts RMS}$$

Now, applying **eq. 4**:

$$G = 20 \log (E_2/E_1) - 10 \log (R_2/R_1) \ [dB] \quad (7)$$

$$G = 20 \log (223.6 \ mV/7.07 \ mV) - 10 \log (50 \text{ ohms}/50 \text{ ohms})$$

$$= 20 \log (31.63) - 0 = +30 \ dB$$

Look carefully at the two results in **eqs. 6** and **7**. If you were told that the gain of this 50-ohm amplifier was 30 dB, would you have to ask "dB-voltage or dB-power?" (I hope not.) As shown in this example, computing the gain by either 10 log of the power ratio or 20 log of the voltage ratio yields exactly the same result in a constant impedance system. So you can see that the all-too-often-asked question has little meaning when the concept of the decibel is properly used.

We will now see how the concept came to be misapplied. Look back at **eq. 4**. This is an exact expression of the decibel and, as explained, reduces to **eq.**

5 in systems of constant impedance. In the early applications of the decibel, power measurements were made in waveguide and coaxial RF systems using various instruments for direct power measurement. The use of these instruments required the signal of interest to be applied directly to the measurement instrument input as is required by the definition of the decibel. Therefore, ratios of measured powers expressed in dB were consistent with the original definition. Then something terrible happened: the performance of electronic instruments improved dramatically. RF voltmeters could now be used to measure actual RMS potentials in systems, rather than only power. Oscilloscopes could provide direct viewing of the voltage waveforms from which RMS values could be computed. And worse yet, these instruments were of such a nature that these potential measurements could be made quite accurately in systems of almost any impedance without the need to break the circuit for direct application of the measured signal to the measuring instrument. In fact, the impedance did not even have to be known to accurately measure potentials, although circuit loading did have to be considered. Well, many of the first applications of these instruments were still in the area of constant impedance systems and it was well known that **eq. 5** applied, and **why**. As several generations of engineers and technicians used these new and ever-improving instruments, the use of **eq. 5** became second nature and its origin (and limitations) slowly became lost and forgotten. Then another terrible thing happened . . . the operational amplifier appeared. These were marvelous devices with staggeringly high voltage gains — perhaps as high as 1,000,000 or even higher! Using such large terms in everyday communication presented a bit of an inconvenience. Then someone, remembering **eq. 5** (at least most of it), said "Wow, we can express this gain in dB as 20 log of the voltage gain." What was omitted was that **eq. 5** applies only in constant impedance systems. Operational amplifiers typically exhibit very high input and very low output impedances. So was yet another misapplication of the decibel born.

To demonstrate the problem associated with this misapplication of the concept, we will examine a few examples. Consider an operational amplifier configured for a voltage gain of unity. Let the amplifier have a 1 Megohm input resistance and a very low output resistance (much smaller than 50 ohms). Also, consider a source with a 50-ohm impedance. Finally, let the load be 50 ohms. If we apply an input signal and measure the input and output voltages we will naturally find them to be the same since the amplifier is configured for a gain of one. Using **eq. 5** and rather ignoring the impedance requirement, we would find the amplifier gain to be 0 dB. Now let's compute the gain in dB as 10 log of the output to input power ratio. The input power is simply the input RMS voltage squared, divided by the input resistance. The output power is given as the output RMS voltage squared divided by the load resistance. However, since the voltage gain is unity, the input and output voltages are equal. Let that voltage be E. Computing the gain from the power ratio:

$$
\begin{aligned}
G &= 10 \log [(E^2/R_L)/(E^2/R_{in})] \\
&= 10 \log (R_{in}/R_L) \\
&= 10 \log (1 \cdot 10^6 \text{ ohms}/50 \text{ ohms}) \\
&= 10 \log (2 \cdot 10^4) = 43 \text{ dB}
\end{aligned}
$$

Well, that presents a bit of a problem. Is the actual gain 0 dB or 43 dB? Let's try still a different measurement by trying to apply the single-point measurement approach of the original definition. Let the 50-ohm load be the input resistance of a 50-ohm power meter. Applying the source to the power meter input, a power, P, is observed. Now, move the power meter to the amplifier output and apply the source to the amplifier input. Since the source is not loaded by the amplifier input (1 megohm $>>$ 50 ohms), the voltage at the amplifier input is twice that measured when the source was terminated with the 50 ohms of the power meter. (This can easily be shown with some simple circuit analysis, but since that's not our purpose here, you'll have to either accept it as true, or prove it for yourself.) The amplifier output voltage is also twice the loaded value of the source, since the amplifier voltage gain is unity and the low output resistance of the amplifier prevents loading by the power meter. Power varies as the square of voltage, so the doubling of the voltage at the power meter input results in an increase in power by a factor of 4. The power reading of the power meter will then be 4P. Applying **eq. 1**:

$$
\begin{aligned}
G &= 10 \log (P_o/P_{in}) \\
&= 10 \log (4P/P) = 6 \text{ dB}
\end{aligned}
$$

This gives us still another choice as to what the gain in dB is for an operational amplifier configured for unity gain. It is either 0 dB, 43 dB, or 6 dB, depending how one makes the measurement.

Now, let's modify the unity gain amplifier circuit configuration slightly by the addition of an input transformer. Let that input transformer match the 1 megohm amplifier input resistance to the 50-ohm source resistance, a turns ratio of 1:141 (remember, transformer impedances vary as the square of the turns ratios). The transformer/amplifier combination now satisfies the constant impedance requirement of the definition of the decibel — the input and output resistance is 50 ohms in a 50-ohm system. The 50-ohm in-

put of the transformer presents the same load as the power meter of the original measurement of the source power. The source power will produce some input voltage. Let that voltage be E. Since the transformer has a 1:141 turns ratio, the input potential to the amplifier is 141E and since the amplifier has unity gain, the output potential is 141E. Since this is a constant impedance system, we may compute the power gain either from the power ratio or the voltage ratio. Therefore, since we have the input and output voltages (in terms of E) we will apply **eq. 5**.

$$G = 20 \log (E_2/E_1)$$
$$= 20 \log (141E/E) = 43 \ dB$$

Now, isn't that an interesting result? This is the same value that was found for the original configuration in the second calculation, which was based on the actual input and output powers. This can somewhat be understood since the transformer can have no power gain. The transformer does provide a proper impedance match to the source so that for any given source potential the maximum amount of power will be coupled, but it does not provide any power gain. The optimum matching simply implies that this is the configuration in which the *available* source power is most effectively coupled. Thus, for some desired output power, this configuration will require the least input voltage, and least available input power. The first unity gain configuration and the transformer/amplifier configuration both have a power gain of 20,000. However, in the first, the voltage gain is unity, so if an output potential of V volts RMS is needed for some desired output power, the input potential must also be V volts. In the transformer/amplifier configuration, the voltage gain is 141, so for an output of V volts an input of V/141 volts is needed. Since each of these configurations is delivering the same power to the load and they have the same power gain, the second makes the more efficient use of the source signal.

Since the gain of this final configuration to which the concept of the decibel may be properly applied is the same as that of the simple unity gain amplifier expressed in dB as 10 log of the actual output to input power ratio, perhaps the gain in dB of the unity voltage gain amplifier should be 43 dB? One can easily see the confusion that can result. This is particularly true since all four of these results could be considered correct depending upon how one chooses to use the concept of the decibel. Considering the first two cases it might have some meaning to use the expressions dB-voltage and dB-power to distinguish between the two measurement techniques. But what do we do with the third measurement? It was also a power measurement and actually more consistent with the actual definition of the dB. Perhaps this value could be called "dB-power-almost-consistent-with-the-definition-of-the-decibel," or dB-PACWTDOTD for short. Certainly that would be a bit ridiculous — but would it be any more absurd than dB-voltage and dB-power? Then there's the modified configuration, to which the concept of the decibel may be properly applied, which yields a power gain that is the same as that of the second calculation. However, this is a different configuration, and its relation to the problem is not immediately obvious. Perhaps the best way to avoid this dilemma is to stick to the actual definition of the decibel. Nevertheless, widespread use of the expression of voltage gain in dB as 20 log of the ratio of the output to input voltages has resulted in this expression becoming a type of alternate definition of the decibel.

One popular use of this expression is the specification of the open loop voltage gain of operational amplifiers. There seems to be a slight trend away from this erroneous specification, however, in favor of the correct units of volts per millivolt, volts per microvolt, etc. Look through some data books with specifications on operational amplifiers and see if you can find these different open loop gain units.

noise analysis — yet another creative misapplication

The use of the decibel to express the voltage gain of an amplifier in a system of non-constant impedances is by far the most common misapplication of the decibel that you're likely to encounter. Yet there's another interesting misapplication that's also quite creative; this is found in the area of noise analysis. The noise power available in a system is directly a function of the system noise bandwidth. Without going into detail, the available noise power, P_n, in a noise bandwidth, BW_n, and absolute temperature, T_O, is given by **eq. 8**.

$$P_n = k \cdot T_O \cdot BW_n \ (watts) \tag{8}$$
$$k = \text{Boltzmann's constant}$$
$$= 1.38 \cdot 10^{-23} \ \text{Joules/degrees Kelvin}$$

Now suppose we have two different noise bandwidths and want to know how much noisier, in dB, the larger is than the smaller. Consider an amplifier system of power gain G with a variable bandpass filter and a suitable power meter tied to the amplifier output. With the bandpass filter set for a narrow bandpass, BW_{n1}, let the power reading be P_{n1}. Then let the filter be adjusted to a wider bandpass, BW_{n2}, with a corresponding power reading of P_{n2}. The relative increase in power with the increased bandwidth expressed in dB is then given as $10 \log (P_{n2}/P_{n1})$, and this expression is an exact application of the definition of the dB. Let this quantity be defined as D. The two powers may

be expressed in the form of eq. 8 above. Then, expanding the expression for D:

$$D = 10 \log (P_{n2}/P_{n1})$$
$$= 10 \log (k \cdot T_o \cdot BW_{n2} \cdot G / k \cdot T_o \cdot BW_{n1} \cdot G) \quad (9)$$
$$= 10 \log (BW_{n2}/BW_{n1}) \text{ (dB)}$$

Well, there you have it — the ratio of two bandwidths expressed in dB is given as 10 log of the bandwidth ratios. Just what does it mean? That's right, *nothing* . . . unless, of course, you know that you're really talking about ratios of noise powers and not simply bandwidths. Feel free to consider some of the possibilities for this misapplication — but please don't take them seriously. Why, one could use eq. 9 (without considering its origin) to express the peaking properties of a bandpass filter in dB as 10 log of the output bandwidth to input bandwidth! Of course in this case we'd most likely want to invert the ratio (i.e., 10 log of the input to output bandwidths) so that we would always have positive values to work with. We could even express quality factor, Q, in dB if we are a little creative. Quality factor of various systems is often expressed as the ratio of the center frequency to the bandwidth. Well, the center frequency could be considered as a bandwidth with the lower frequency of 0 Hz. Then Q would be nothing more than a ratio of two bandwidths and misusing eq. 9, we could express Q in dB (again, we would wish to invert the ratio to have positive dB values of Q since negative values might be confusing).

If these suggestions seem quite outrageous, perhaps it's because they haven't been seen before. These are actually as correct as expressing the open loop voltage gain of an operational amplifier in dB, but we've come to know *that* expression because of its widespread use, and so as a result, it doesn't look particularly strange. But this doesn't make it inevitably correct. There is some obvious confusion because in the absence of uncertainty, the voltage/power question would need never be asked. If you stick to using the concept of the decibel only where it applies, you'll have no problem. In cases where you must work from someone else's error, you'll just have to try to figure out what was meant. This does not mean that we would specify the gain of the unity gain amplifier in the example above as 43 dB. That voltage gain is 1 volt/volt or simply a voltage gain of 1. The power gain of that configuration is 20,000. That is where we arrived at the 43 dB figure; but as pointed out, the concept of the decibel does not apply because the system is not of constant impedance. The power gain would simply be stated as 20,000 watts/watt, 20 watts/milliwatt, or simply a power gain of 20,000.

As the examples above have shown, misapplication of the concept of the decibel has led to considerable confusion as to what is actually implied by the expression of a quantity in dB. Hopefully this article has served to clarify your understanding of the concept, and you can now correctly apply the concept. Furthermore, when you see quantities expressed in dB, you'll be able to tell whether or not they're proper expressions. In cases where they're not, you should have a better understanding of the concept to aid you in trying to comprehend the original intention. In any event, if you keep the definition of the decibel clearly in mind and always apply it properly, then your expressions, at least should always be correct and should be understood by anyone who shares your understanding of the decibel. When you're asked, "Is that dB-voltage or dB-power?", you'll not only be able to show why that question has little meaning, but also silently revel in your mastery of the concept. Always return to the basic definition of a notation or concept and apply it accordingly. Also, look those basic definitions up in the literature. Don't simply take some *expert's* opinion as being correct. Perhaps you should apply that advice to the information in this article!

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ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

the G5RV explained

While it may be premature to speculate, Spring can't be far away. Now's the time to think about that new antenna you're going to put up as soon as milder weather rolls around.

Some Amateurs in the United States have been bemused, even confused, by the short, cryptic description of an antenna used by many European Amateurs. Described as simply a "G5RV," the antenna must work, judging from some of the powerful signals that come "across the pond" from stations using this sky-wire.

The antenna design is named after its designer, Louis Varney, G5RV, an old-timer — (licensed as 2ARU in 1928, and as G5RV in 1929) — still very active on the bands. Louis has used the antenna from many of his overseas DX locations over the past few decades. It enjoys worldwide popularity because it's a good, inexpensive multiband antenna that works very well.[1]

The grandfather of the G5RV was an "all-band" antenna first described in Amateur literature by Arthur Collins, WØCXX, then President of Collins Radio Company. Sold as a kit in 1933, it never attained widespread popularity because it was both expensive and difficult to install.

The Collins antenna consisted of a 3/2-wavelength long 20-meter, center-fed wire with an impedance transformer that provided a satisfactory match to the open-wire line and tuned tank circuit rigs of the pre-war period. The transformer was a linear affair made of copper tubing that was difficult to suspend from the center of the antenna. Art's transformer was quickly forgotten, but the idea of the inexpensive, effective 3/2-wave antenna remained (**fig. 1**). As shown in **fig. 2**, it has an interesting field pattern and a radiation resistance value of about 95 ohms at the center feedpoint. Its power gain over a dipole is about 1 dB.

Antenna dimensions for the higher frequency bands are shown in **table 1**. The antenna can be easily matched to 50-ohm coaxial line by means of a quarter-wave section of 75-ohm coaxial line. Sufficient line isolation can be obtained by wrapping a portion of the 75-ohm line into a simple four-turn RF choke coil directly beneath the antenna feedpoint.

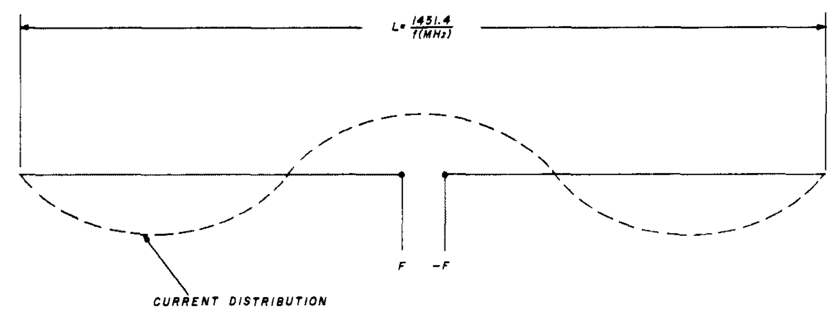

fig. 1. The 3/2-wave center-fed antenna. Feedpoint resistance at resonance is about 95 ohms.

This simple antenna is recommended as a general coverage, single-band antenna for the Amateur bands between 10 and 30 MHz.

a practical G5RV multiband antenna

Louis Varney, G5RV, devised a 1/2-wave matching section that functions as a 1:1 transformer for a 14-MHz 3/2-wave dipole, enabling the coaxial line to "see" a close match on this band (**fig. 3**). On other high frequency Amateur bands, the transformer section acts as a portion of the antenna, folded back upon itself, to provide a reasonable value of feedpoint impedance on all bands between 3.5 and 29.7 MHz. Even though the antenna termination may be reactive, a good antenna tuning unit will match the system to a 50-ohm transmitter. This scheme will satisfy the rather stringent load conditions required by many of the solid-state transceivers employing a "fail-safe" ALC circuit that senses the SWR on the transmission line and reduces amplifier input to protect the output transistors of the transceiver from damage.

It is tempting to substitute a balun for the antenna tuner to make a "no adjustment," multiband antenna. This is an unsatisfactory solution. Ferrite-core baluns are not effective with reactive loads or loads presenting a high value of SWR, and cannot compensate for the reactive load presented by the G5RV antenna on most Amateur bands. The tuners listed in references 2 and 3, however, will do the job properly.

the linear transformer for the G5RV

The heart of the G5RV antenna is

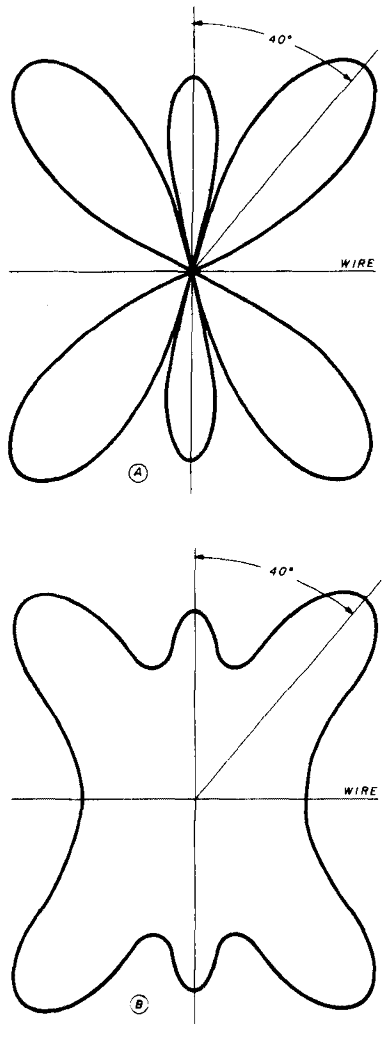

fig. 2. Horizontal field pattern for 3/2-wave antenna: (A) ideal pattern; (B) over imperfect ground.

table 1. Dimensions of flat-top and coaxial transformer for 3/2-wavelength, center-fed antenna.

		L = *		1/4-wave transformer (RG-11/U or RG-59/U	
band	f(MHz)	feet	(meters)	feet	(meters)
30	10.12	143.40	(43.70)	16.04	(4.89)
20	14.15	102.60	(31.30)	11.46	(3.49)
17	18.11	80.14	(24.43)	8.97	(2.73)
15	21.22	68.40	(20.85)	7.66	(2.33)
12	24.94	58.20	(17.74)	6.51	(1.98)
10	28.60	50.75	(15.47)	5.68	(1.73)

$$*L = \frac{1451.4}{f(MHz)}$$

the 14-MHz 1/2-wave transformer placed at the feedpoint of the antenna. Line impedance is not important. Ideally, an open-wire line is best, but is difficult to build and spacers are hard to come by. A good substitute is TV-type "ladder line" that will function with power levels up to 250 watts or so. On occasion, transmitting-type ladder line that will work very well can be found.

Alternatively, 300-ohm line having "windows" punched in the dielectric can be used, but this material is not difficult to obtain. While 300-ohm TV ribbon line will work, the transformer section can be detuned by rain or snow, making antenna tuning erratic in wet weather.

Regardless of construction, the transformer section should drop down vertically beneath the antenna for 20 feet (6 meters) or so before it is brought away at an angle. This will keep undesired coupling between line and antenna at a minimum.

The G5RV can be installed as an inverted-V antenna and still perform successfully.

the 160-meter compact dipole

Have you heard the DX coming through on the 160-meter band? Would you like to work some of it? A great idea, but a lot of would-be "top band" DXers pause when they consider that a 1/2-wave dipole antenna is about 246 feet (75 meters) long when cut for the midpoint of the band. And while a vertical antenna would be appropriate, it requires a good ground connection and a system of buried radials.

An effective alternative is a short, coil-loaded dipole antenna. By reducing the dipole to half size, about 130 feet (40 meters), the antenna becomes more feasible for the ham who lives on a medium-sized lot. A loaded antenna can be just about any length, however, if a compromise between length, efficiency and bandwidth can be accepted. Bandwidth and efficiency drop

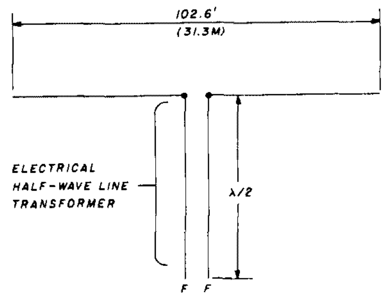

fig. 3. The G5RV antenna for 14 MHz. A parallel-wire transformer 1/2-wavelength long acts as a 1:1 matching device on 14 MHz and as a folded portion of the antenna on other bands. A tuner is required at F-F to match the system to a coaxial feedline. The velocity factor of the line must be used to determine physical line length. On 80 meters the antenna is a foreshortened dipole. On 40 meters the antenna is a folded "two 1/2-waves in phase." On 18, 24 and 28 MHz the antenna functions as a long wire, center-fed.

sharply when the loaded dipole is much less than 1/4-wavelength long.

The bandwidth of a full-size 160-meter dipole mounted close to the ground (40 to 60 feet — or 12 to 18 meters — high) is quite narrow — only about 150 kHz between the 2:1 SWR points on the passband. Shortening the dipole and loading it to resonance sharpens the passband. The antenna design shown in **fig. 4** has a passband of about 50 kHz between the 2:1 SWR points. That's the penalty you pay to get a compact antenna on 160

table 2. Coverage of the entire 160-meter band requires changing dimensions and component values.

design frequency (MHz)	overall length (S) feet	(meters)	length center-to-coil (D) feet	(meters)	loading coil (L_1) μH	turns (L_1)	matching coil (L_2) μH	turns (L_2)
1.80	130.0	(39.6)	32.5	(9.9)	91.9	38.9	3.9	11.6
1.85	126.5	(38.6)	31.6	(9.6)	89.2	38.0	3.8	11.3
1.90	123.2	(37.6)	30.8	(9.3)	86.5	37.2	3.7	11.1
1.95	120.0	(36.6)	30.0	(9.1)	84.0	36.4	3.6	11.0
2.00	117.0	(35.66)	29.3	(8.9)	81.6	35.7	3.5	10.8

Notes:

Dimensions rounded to nearest tenth, metric dimensions in ( )
Coil L_1 diameter = 3 inches (7.6 cm) use No. 14 wire, close-spaced
Coil L_1 length = approximately 2.5 inches (6.3 cm)
Coil L_2 diameter = 1 inch (2.54 cm), use No. 18 wire, close-spaced
Operating bandwidth = 50 kHz between 2:1 SWR points
Adjust tip sections for antenna resonance
Adjust coil L_2 for best match at resonant frequency

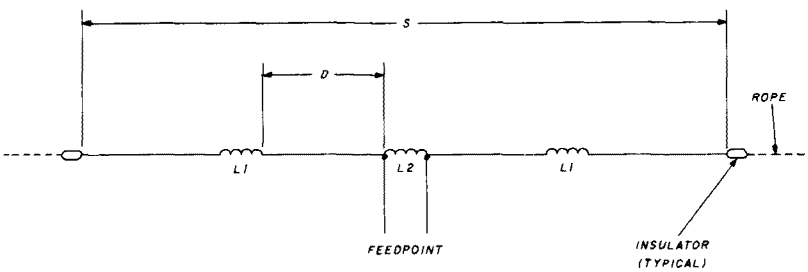

fig. 4. The 160-meter compact dipole (see table 2).

meters! You can make the dipole shorter with larger loading coils, but your operating passband will shrink until it becomes impractically narrow.

simplifying the design

The chart shown in **table 2** reveals a number of interesting points. Overall antenna length varies from 130 feet (39.6 meters) at the low end of the band to 117 feet (35.6 meters) at the top end. That's a difference of 13 feet (4 meters). The length of the center-to-coil distance also changes appreciably (from 32.5 feet to 29.3 feet). The loading coil (L1) inductances change only slightly, as the number of turns changes only from 38.9 to 35.7. And the matching coil at the center of the antenna changes hardly at all.

It seems to me that things could be simplified by using the center-band design (1.90 MHz) and then varying the resonant frequency of the whole antenna by merely changing the length of the tip sections. Leave all the other dimensions alone. Fold-back tip sections can be wrapped around the antenna wire and then unwrapped and left to hang down when operation is desired over a lower frequency range.

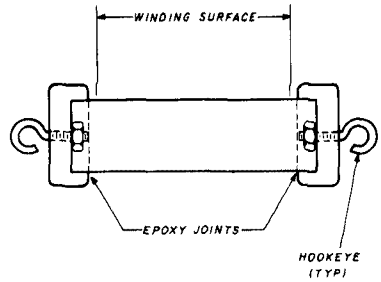

fig. 5. Loading coil can be made up of PVC-type plastic pipe and end caps. Hookeyes permit connection to the antenna wire.

adjusting the antenna

Accepting the 1.90 MHz dimensions as par, then, what's to be done? The antenna is built, erected in place, and lowered so that a dip-meter can be coupled to the matching coil, L2. The end sections of the antenna are trimmed equally until resonance is established at 1.90 MHz, or at any other point you decide is your "pet" operating frequency. (The feedline is removed for this test.)

Once antenna resonance is determined, matching coil L2 is adjusted, a quarter-turn at a time, for the lowest SWR indication on the feedline at the frequency of antenna resonance. The antenna must be in the final operating position when this is done.

building the antenna

The loading and matching coils can be easily made up using PVC tubing, as shown in **fig. 5**. If the coils are given a good coat of acrylic spray, they'll be weatherproof. Some detuning may be noticed in damp or wet weather, or if snow clings to the coils.

An alternative for Amateurs living in mild climates is to make the coils of prefabricated coil stock. The latest Barker & Williamson catalog lists show

both "Airdux" and "Miniductor" coils.* I understand the "Airdux" coils can be specially ordered with LEXAN® insulation, which is impervious to ultra-violet rays. This means that the coil support strips will not disintegrate after being exposed to sunlight over a period of time.

the easy way out

If *you don't* want to build the compact 160-meter dipole yourself, you can purchase either the loading coils (Model LC-1) or the complete antenna (Model AES-160) from Barker & Williamson. Keep in mind, however, that this antenna is shorter than the one described earlier in this article, and while it works just as well, it may have a smaller passband because of its shorter 96-foot length. Take your choice.

heavy-duty equipment

Do you need mica transmitting-type capacitors? One possible source is Milton Levy, W5QJT, Apartment 303, 350 North Resler Drive, El Paso, Texas 79912. Milt has a large collection of capacitors and vintage radio tubes for sale at modest prices. (By the way, have you priced receiving tubes lately? The new Newark Electronics catalog lists a 6SJ7 at $29.46 and a 6SN7 at $16.82.)†

lightning protection

Many VHF/UHF repeater antennas are mounted to the side of a metal tower and grounded to it. Even so, the antenna and equipment can be damaged by the electric field of a nearby lightning stroke. A simple and effective way to protect the side-mounted antenna is shown in **fig. 6**. A "lightning rod" is mounted to the tower just above the repeater antenna. The horizontal metal rod, a few inches longer than the spacing between the antenna and tower, is placed 6 inches or more above the tip of the antenna. The lightning rod will not affect antenna performance, but it will help to protect the antenna and the equipment attached to it during a nearby storm.

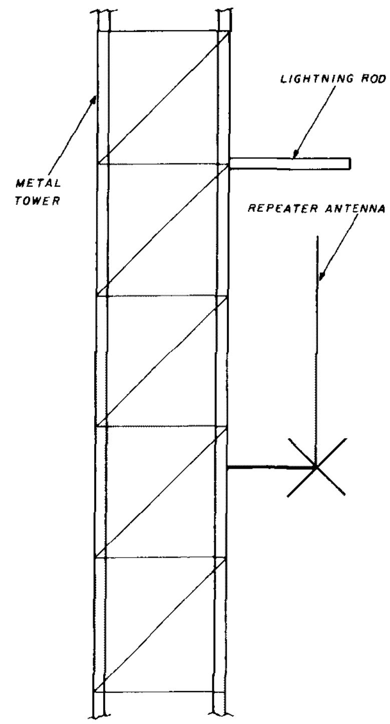

fig. 6. Lightning rod will protect repeater antenna mounted on side of metal, grounded tower.

reprint available

I have a limited number of reprints of the article "A High Power 2-Meter Amplifier Using the New 3CX800A7" from the April, 1984, issue of *QST*. Address your request to me, c/o EIMAC, 301 Industrial Way, San Carlos, California 94070. (Include two first-class stamps or two IRCs.) Thanks to the American Radio Relay League for permission to reprint.

*Barker & Williamson, 10 Canal Street, Bristol, Pennsylvania 19007.

†Newark Electronics, 500 North Pulaski Road, Chicago, Illinois 60624.

references

1. Louis Varney, G5RV, "G5RV Multiband Antenna . . . Up-to-date," *Radio Communications*, July, 1984, page 572.
2. "Band Switched Link Coupler," *The ARRL Antenna Book*, 14th edition, 1982, page 4-4 and 4-5.
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ham radio

Build a compact, powerful, portable unit for under $300

mobile solid-state kilowatt HF linear amplifier

Not too long ago my mobile kilowatt linear amplifier barely fit inside the trunk of my car and raised havoc with the electrical system each time a heavy load was applied. Because the efficiency of the old tube-type linears was low, the current drawn from the battery was high. Sometimes it even caused the car to stall. But progress in transistor technology has brought the mobile HF linear within reach of most Amateurs. If you're handy with a drill, a screwdriver, and a soldering iron, you're an ideal candidate for this project. No special skills are required beyond the desire to own a high-power mobile linear and the willingness to do some assembly work.

The efficiency of the amplifier is as high as 70 percent, depending on the condition of your battery and the accuracy of your antenna match. With a DC input of 830 watts, output power on 75 meters is as high as 580 watts. The total cost for the entire circuit — assuming a good supply of junk box parts is available — should be under $300. If you can get the transistors for free, the amplifier shouldn't cost you more than $150.

table 1. Test data on amplifiers matched for low end of HF band (2-7 MHz)

band (meters)	input drive (watts)	VSWR in	CW power out (watts)
160	35	1.2:1	500
75	30	1.2:1	500
40	45	1.5:1	450
20	50	2.0:1	450
15	50	2.5:1	350

amplifier circuit description

The linear amplifier consists of four amplifier modules capable of PEP power output levels in excess of 150 watts each (see **fig. 1**). Each module uses a pair of 75 watt RF transistors in a push-pull configuration for maximum efficiency and lowest possible distortion (see **fig. 2**). Both input and output impedances of each transistor module are preset at 200 ohms, which makes combining relatively easy. The input and output divider/combiner provides an ideal match with very low loss from the modules to the input/output connectors (see **fig. 3**).

A drive requirement of 50 watts was selected to allow the amplifier to be used with the standard exciters at reduced output. The actual gain of the amplifier depends on the frequency used and the transistors selected for the project. I had good results with the TRW PT9784 type. The transistors are slightly forward biased for good linear performance and maximum efficiency. No fancy biasing circuits are needed; just a few inexpensive resistors and a diode are sufficient. I found that temperature stability was not

By Frank Kalmus, WA7SPR, 7016 NE 138th Street, Kirkland, Washington 98034

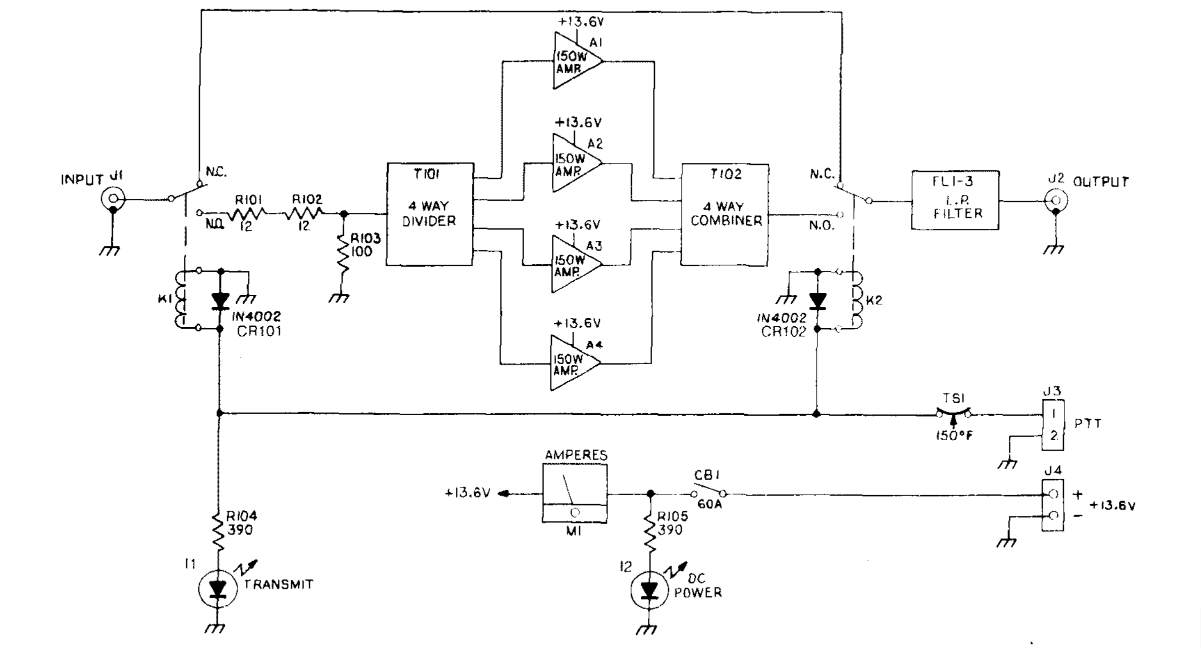
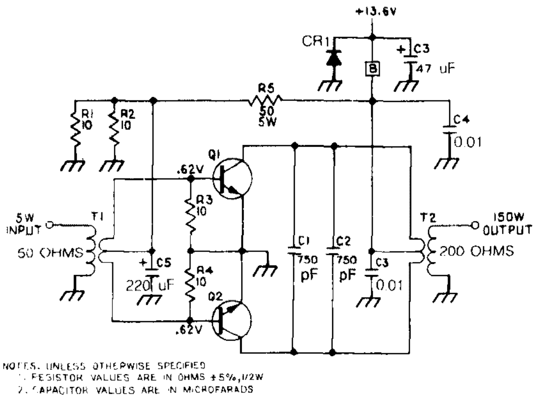


fig. 1. Mobile kilowatt block diagram.

a problem even when the unit was operated on CW for long periods of time. The two relays that form the T/R switching are keyed by the PTT line, which is wired for 12 volts input from the transceiver in transmit mode. If your exciter doesn't provide you with 12 VDC when keying the microphone, you can rewire the PTT relay circuit so it will key when a ground is provided from the exciter.

The *T/R switch* provides automatic switching between the receiver and the amplifier during transmit and receive modes as well as during exciter "barefoot" operation (see **fig. 4**). When the circuit breaker is switched off, the amplifier is bypassed. With the DC power breaker in the ON position, the linear is switched on every time you key the microphone; you are then transmitting at a kilowatt without having to worry about high voltages or battery drain.



T1 input transformer

quantity	***description***
2	***ferrite bead, 2643006301***
2	***PC board ends***
2	***brass rod, 3/4 inch (small rod)***

T2 output transformer

2	***ferrite bead, Stackpole No. 57-2857***
2	***PC board ends***
2	***brass rod, 1-7/16 inch (large rod)***

fig. 2A. Schematic of basic building block: the 150-watt amplifier.

item	*description*	*quantity*
B	*bead, shield*	*3*
C1,C2	*capacitor, SM 750 pF*	*2*
C3	*capacitor, 0.01 μF, 50 V*	*1*
C4	*capacitor, 47 μF, 35 V*	*1*
C5	*capacitor, 220 μF, 35 V*	*1*
C6	*capacitor, 0.01 μF, 1 kV*	*1*
CR1	*diode, MR500*	*1*
Q1,Q2	*transistor, RF, 75 watts TRW PT9784 or Motorola MRF454 and MRF412, MRF458*	*2*
R1-R4	*resistor, 10 ohm, 1/2 watt*	*4*
R5	*resistor, 50 ohm, 5 watts*	*1*
	PC board	*1*
T1	*transformer, input*	*1*
T2	*transformer, output*	*1*

fig. 2B. Parts list for 150-watt RF amplifier (four required).

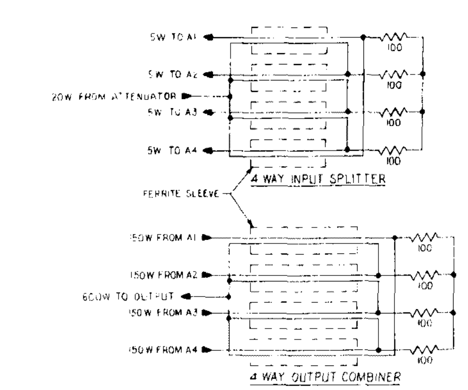


fig. 3A. Interconnection diagram for 4-way input splitter and output combiner.

description	*quantity*
PC board	*2*
bead, Stackpole No. 57-2857	*4*
standoff, Sealectro, 023-5702-000-709	*9*
Insulator, H.H. Smith No. 2604	*1*
resistor, 220 ohm, 2 watts	*8*
No. 16 buss — 4-1/2 inch	*8*
sleeving — 2.9 inch (teflon)	*8*

fig. 3B. Parts list for 4-way input splitter and output combiner.

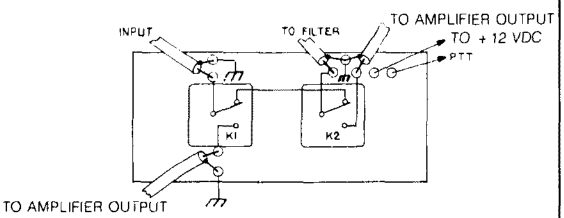
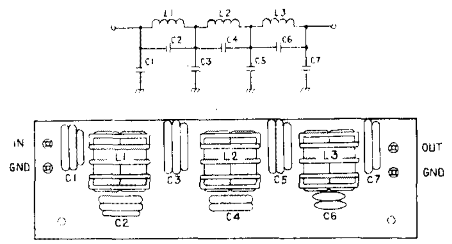


fig. 4. Transmit/receive, switch connections.

The *input attenuator* consists of a resistive power divider that incorporates high frequency compensation to achieve bandpass response flatness. The resistors used are of the "non-inductive" type and are rated at 25 watts.

A 60-ampere panel meter monitors the DC current drain from the battery. (This was achieved using a 40-amp meter. The dial reading is multiplied by 1.5.) On CW it is possible to get readings as high as 60 amperes. Though difficult to read on the meter, PEP currents on SSB are as high as 75 amperes. The average SSB power consumption is less than 200 watts, or about 15-18 amperes DC. The average automobile battery and electrical system should not have any problem powering this amplifier. It is,



band (meters)	*L1*	*L2*	*L3*	*C1*	*C2*	*C3*	*C4*	*C5*	*C6*	*C7*
75/80	*2.0*	*1.6*	*1.6*	*458*	*78*	*848*	*370*	*760*	*296*	*290*
40	*1.2*	*0.9*	*0.9*	*282*	*47*	*526*	*229*	*470*	*180*	*180*
15/20	*0.4*	*0.4*	*0.4*	*94*	*15*	*182*	*82*	*164*	*68*	*68*

fig. 5A. Low-pass filter schematic, parts placement diagram, and component values.

item	*description*	*quantity*
C1	*capacitor, 390 pF, ceramic, 2 kV, 5 percent*	*1*
C1	*capacitor, 88 pF, ceramic, 3 kV, 5 percent*	*1*
C2	*capacitor, 39 pF, ceramic, 3 kV, 5 percent*	*2*
C3	*capacitor, 390 pF, ceramic, 2 kV, 5 percent*	*2*
C3	*capacitor, 88 pF, ceramic, 3 kV, 5 percent*	*1*
C4	*capacitor, 270 pF, ceramic, 2 kV, 5 percent*	*1*
C4	*capacitor, 100 pF, ceramic, 4 kV, 5 percent*	*1*
C5	*capacitor, 390 pF, ceramic, 2 kV, 5 percent*	*1*
C5	*capacitor, 270 pF, ceramic, 2 kV, 5 percent*	*1*
C5	*capacitor, 100 pF, ceramic, 4 kV, 5 percent*	*1*
C6	*capacitor, 120 pF, ceramic, 3 kV, 5 percent*	*2*
C5	*capacitor, 56 pF, ceramic, 3 kV, 5 percent*	*1*
C7	*capacitor, 120 pF, ceramic, 8 kV, 5 percent*	*2*
C7	*capacitor, 50 pF, ceramic, 3 kV, 5 percent*	*1*
L1	*inductor, 2 μH*	*1*
L2,L3	*inductor, 1.8 μH*	*2*
	PC board	*1*

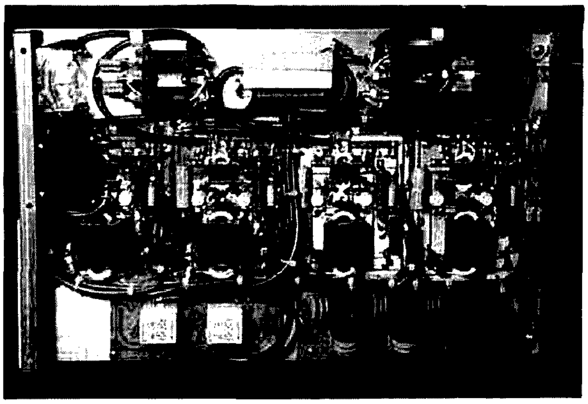
fig. 5B. Parts list for FL1 filter subassembly (75 and 80 meters). Note: In this and other parts lists, repetition of the line designation signifies adding these components to make *one* effective component. For example, under C1, 390 and 68 pF capacitors are combined to achieve 458 pF.

item	*description*	*quantity*
C1	*capacitor, 100 pF, ceramic, 4 kV, 5 percent*	*2*
C1	*capacitor, 82 pF, ceramic, 5 kV, 5 percent*	*1*
C2	*capacitor, 47 pF, ceramic, 3 kV, 5 percent*	*1*
C3	*capacitor, 270 pF, ceramic, 2 kV, 5 percent*	*1*
C3	*capacitor, 200 pF, ceramic, 3 kV, 5 percent*	*1*
C3	*capacitor, 56 pF, ceramic, 3 kV, 5 percent*	*1*
C4	*capacitor, 100 pF, ceramic, 3 kV, 5 percent*	*1*
C4	*capacitor, 82 pF, ceramic, 5 kV, 5 percent*	*1*
C4	*capacitor, 47 pF, ceramic, 3 kV, 5 percent*	*1*
C5	*capacitor, 270 pF, ceramic, 2 kV, 5 percent*	*1*
C5	*capacitor, 200 pF, ceramic, 3 kV, 5 percent*	*1*
C6	*capacitor, 68 pF, ceramic, 3 kV, 5 percent*	*1*
C6	*capacitor, 56 pF, ceramic, 3 kV, 5 percent*	*2*
C7	*capacitor, 68 pF, ceramic, 3 kV, 5 percent*	*1*
C7	*capacitor, 56 pF, ceramic, 3 kV, 5 percent*	*2*
L1	*inductor, 1.2 μH*	*1*
L2,L3	*inductor, 0.9 μH*	*2*
	PC board	*1*

fig. 5C. FL2 filter subassembly (40 meters). See note in fig. 5B.

item	description	quantity
C1	capacitor 47, pF, ceramic, 3 kV, 5 percent	2
C2	capacitor, 15 pF, ceramic, 5 kV, 5 percent	1
C3	capacitor, 82 pF, ceramic, 5 kV, 5 percent	1
C3	capacitor, 100 pF, ceramic, 4 kV, 5 percent	1
C4	capacitor, 02 pF, ceramic, 5 kV, 5 percent	1
C5	capacitor, 39 pF, ceramic, 3 kV, 5 percent	1
C5	capacitor, 56 pF, ceramic, 3 kV, 5 percent	1
C5	capacitor, 68 pF, ceramic, 3 kV, 5 percent	1
C6	capacitor, 68 pF, ceramic, 3 kV, 5 percent	1
C7	capacitor, 68 pF, ceramic, 3 kV, 5 percent	1
L1	inductor, 0.4 μH	1
L2,L3	inductor, 0.37 μH	2
	PC board	1

fig. 5D. FL3 filter subassembly (15-20 meters). See note in fig. 5B.



All components are clearly visible in bottom view of kilowatt linear amplifier.

however, suggested that you use at least No. 8 wire for the DC power cable.

filtering cooling and testing

RF filtering is necessary to meet FCC requirements of minimum 40 dB harmonic rejection (see **fig. 5**). A seven-pole elliptical filter is used in the 75-meter unit. Typical harmonic performance was better than 50 dB down from the main carrier.

Cooling was found to be no problem on SSB. Average voice conditions will not cause dangerous overheating if the amplifier is mounted in a well-ventilated area. But just in case, a temperature sensor switch inserted in the T/R relay circuit will shut the amplifier down if the temperature exceeds a preset level of approximately 150 degrees F, thus preventing serious damage to the equipment.

After all the assembly and wiring are done, some preliminary testing is required before applying power to the amplifier (see **fig. 6** for overall assembly parts layout). Using an ohmmeter, connect the black lead to ground and the red lead to the 12-volt line to make sure there are no shorts. Connect the red lead to each collector of all eight transistors to check for short circuits. Measure approximately 2.5 ohms to ground at each base and 12.5 ohms to ground at each collector. Measure continuity from the center pin of the input to the center pin of the output connector. However, neither one of the RF connector center pins should read short to ground.

Because this mobile amplifier was designed for

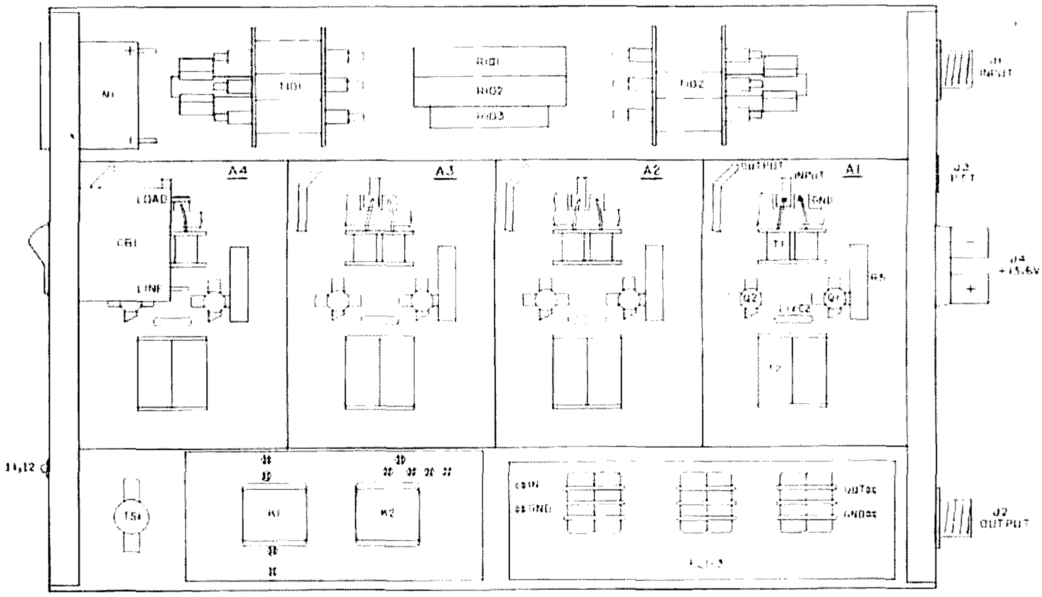


fig. 6A. Overall assembly and parts location guide.

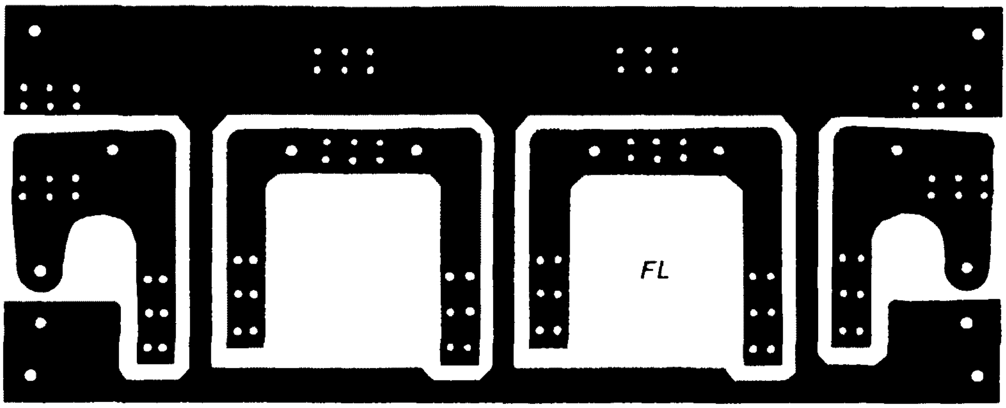


fig. 7A. Filter printed circuit board.

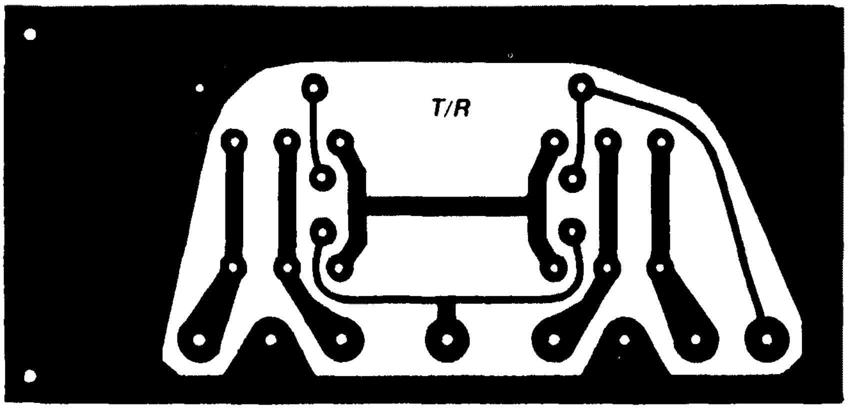


fig. 7B. T/R switching printed circuit board.

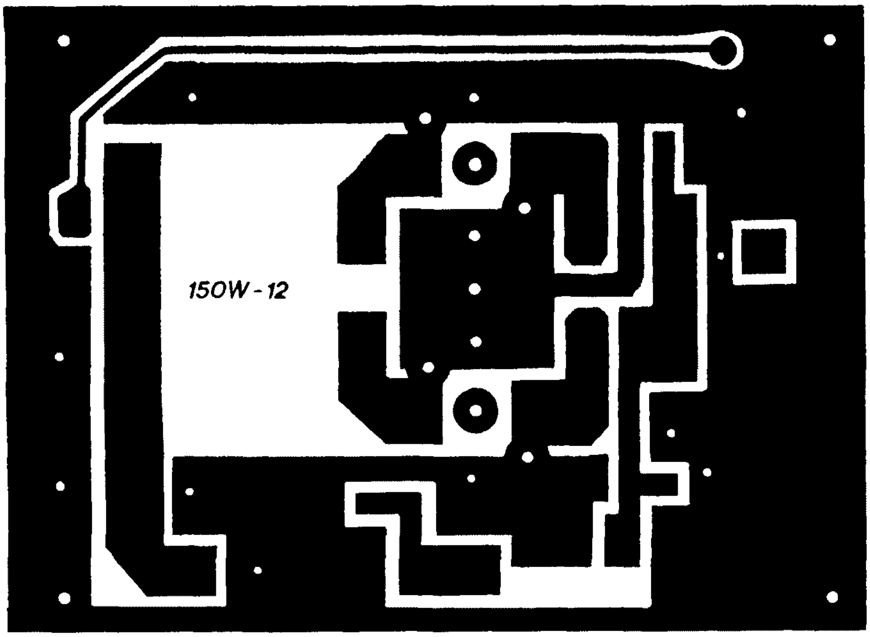


fig. 7C. 150-watt amplifier printed circuit board.

single band operation, a single low-pass filter was installed. If you want to use yours on more than one band, you can build the three filters described in the parts list and use them externally. If you decide to do so, be careful not to operate the amplifier on the "wrong" band for the filter used; doing so can cause permanent damage to the amplifier. It will also be necessary to delete the built-in filter unless you install the 15-20 meter filter internally and use the two other filters externally. The built-in filter will automatically pass both other bands if left in permanently.

item	description	quantity
cabinet	panel, front	1
and	panel, rear	1
heatsink	cover, bottom	1
enclosure	bracket, mounting	1
	heatsink, 15 inches, Thermalloy No. 6130	1
A1-A4	amplifier, RF, 150 watt	4
CB1	circuit breaker, 60 amps, Airpax No. UPGX6-1-52-503	1
CR101,CR102	1N4002 diode	2
FL1-3	filter, low-pass	3
I1	lamp, LED red, Monsanto No. MV5053	1
I2	lamp, LED green, XC556-G	1
J1,J2	connector, Amphenol No. SO238	2
J3	connector, PTT, Cinch No. S-302-AB	1
J4	connector, DC power, H.H. Smith No. 269RB	1
K1,K2	transmit/receive relay, Guardian No. 1365PC-2C-12 VDC	2
M1	meter, 0-60 amperes	1
R101, R102	resistor, non-inductive, 12 ohm, 25 watt	2
R103	resistor, non-inductive, 100 ohm, 10 watt	1
R104,R105	resistor, 390 ohm, 1/2 watt	2
T101,T102	divider/combiner	2
	T/R PC board	1
	cables, various, 50 ohm	16
	lockwasher, No. 4	95
	screw, 4/40 × 3/8-inch round	24
	screw, 4/40 × 1/2-inch round	7
	screw, 4/40 × 5/16-inch round	16
	screw, 6/32 × 5/16-inch flat	10
	screw, 6/32 × 5/16-inch round	6
	spacer, 1/4-inch	7
	standoff, teflon, Useco No. 1456	3
	washer, No. 4 flat	48
TS1	thermal switch, Therm-o-disc No. CLR 160°	1

fig. 6B. Parts list for overall assembly.

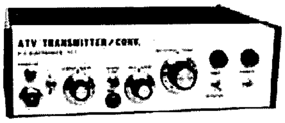
conclusion

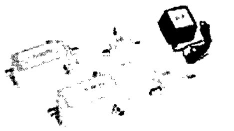
Although a complete and detailed parts list, including schematics and PC board layouts (see **fig. 7A, B, C**), has been provided, this article is intended for use by the licensed Radio Amateur only. Additional information can be provided by the author at the reader's request by mail only, provided the request is accompanied by a copy of a valid Amateur Radio license and a SASE. No other inquiries will be answered.

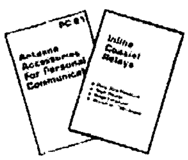
This amplifier was designed to be used on 160/80/75/40 and 20 MHz. Any attempt to extend the frequency range will most likely result in the destruction of the RF power transistors.

Parts for this project are available from the author at $495.00 for all parts included in the parts list, or $339.00 for all listed parts except transistors. Only one filter (for 15/20 meters) is included in these packages. Contact Frank Kalmus, WA7SPR, 7016 NE 138th Street, Kirkland, Washington 98034, for information.

ham radio





DX FORECASTER

Garth Stonehocker, KØRYW

more on ionosphere matching

A method of tailoring an antenna system for maximum signal at a DX location was presented in last month's column. A short computer program for determining the angle of the maximum signal lobes from horizontal antennas, for specific antenna heights above ground, was included.

Correlating this take-off angle with distance requires a few calculations, however. For an approximation, first divide the distance between the transmitting and receiving location (determined from a map with mileage markers or a globe) into equal interval hops. Then, using this hop distance, refer to **fig. 1** to find the required departure (or take-off) angle.

But how can you determine which of the several ionospheric heights provided in **fig. 1** should be used? Simply refer to **table 1**, which indicates representative mid-latitude heights appropriate for use at various times of the day. Using **fig. 1** and **table 1** for the computer program modifications shown in **fig. 2**, you have sufficient information to determine the antenna height that will enhance your DX performance.

To use **table 1**, add the increment (cumulatively) for each month between the months listed or subtract the increment after June. For example, the height of the ionosphere in March at noon local time is: 235 + (33•2) = 301 km. For those who have a computer, the MINIMUF 3.5 program can be used to obtain several useful operating parameters, including take-off angle (departure angle) for the path of interest.[1] Modify the program by inserting the new lines indicated in **fig. 2**.

This modification provides an hourly indication of MUF with great circle distance in kilometers, azimuthal bearings to station and take-off angle in degrees for the path.

last-minute forecast

The second week of the month is expected to favor DX on the high bands, with 10 to 30 meter performance correlating with the beginning of a more energetic flux and sunspot period for the year. Transequatorial propagation, enhanced by any geomagnetic field disturbances toward the end of the week, should provide the best openings of the month.

Listen to WWV at 18 minutes after the hour for high values of the A and K indices. The last week will probably be best for low-band operation although there is a probability that the geomagnetic field will also be disturbed then. Work the stations while you can before the static builds over the coming months.

No significant *meteor* showers are scheduled to appear in February. A full moon will occur on the 5th, with its perigee on the 8th.

band-by-band summary

Ten meters will be open (local time) to the southeast for a short period before noon, to the south at noon, and to the southwest in the afternoon. The openings will be longer and more frequent when the solar flux is at its 27-day cycle maximum.

Fifteen and twenty meters, almost always open to some part of the world, will be the main daytime DX bands. Twenty should stay open on long southern paths into the night, while 15 will drop out in the late afternoon. Operate 15 first and move down to 20 meters. DX is 5000 to 7000 miles (8000 to 11,200 km) on these bands and one-long-hop transequatorial propagation is also possible even more often than on 10 meters. In addition transequa-

table 1. Approximate diurnal height of ionosphere during the year.

local time:	midnight	6 A.M.	noon	6 P.M.
height (km):	all 283	265-290	235-400	225-310
increment:		5	33	30
month:		January - June	January - June	Equinox - June

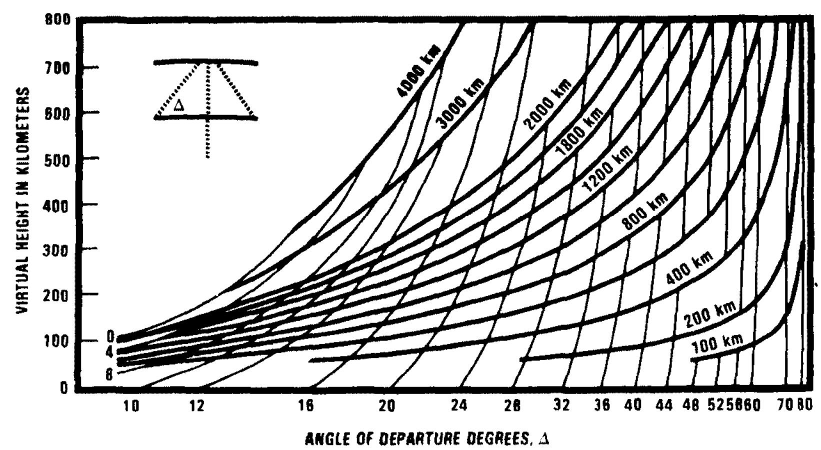


fig. 1. Transmission curves showing vertical angles.

FEBRUARY

WESTERN USA

GMT	2300	2200	2100	2000	1900	1800	1700	1600	1500	1400	1300	1200	1100	1000	0900	0800	0700	0600	0500	0400	0300	0200	0100	0000	
PST	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	
ASIA FAR EAST	20	30	30	40	40	40*	40	40*	30	30	30	30	30	30	30	20	20	20	20	20	20	20	20	20	N
EUROPE	20	20	20	20	20	20	20	20	30	30	30	40	40	40	40	30	30	30	30	30	30	30	20	20	NE
S. AFRICA	15	10	10	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	E
S. AMERICA	15*	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	15	10	10	10	10	10	10	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	15	15	S
NEW ZEALAND	10	10	10	15	15	15	15	15	20	20	20	20	20	20	20	20	20	15	15	15	15	10	10	10	SW
OCEANIA AUSTRALIA	10	10	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	15	10	10	10	10	10	W
JAPAN	20	20	20	20	30*	30	30	40*	40*	30	30	30	30	30	30	30	20	20	20	20	20	20	20	20	NW

MID USA

MST	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	
ASIA FAR EAST	30	30	30	30	20	20	20	20	20	20	20	20	30	30	30	40	40	40	40	40	30	30	30	30	N
EUROPE	20	20	20	20	20	20	20	20	20	30	30	30	40	40	40	40	40	30	30	30	30	30	20	20	NE
S. AFRICA	15	10	10	10	10	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	E
S. AMERICA	15	15	10	10	10	10	10	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	S
NEW ZEALAND	10	10	10	10	15	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	10	10	10	SW
OCEANIA AUSTRALIA	10	10	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	10	10	10	10	W
JAPAN	20	20	20	20	30	30	30	30	30	40*	40*	30	30	30	30	30	30	20	20	20	20	20	20	20	NW
CST	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	

EASTERN USA

EST	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	
ASIA FAR EAST	30	30	30	20	20	20	20	20	20	20	20	20	20	30	30	30	40	40*	40	40	30	30	30	30	N
EUROPE	20	20	20	20	20	20	20	20	20	30	30	30	40	40	40	40	40	30	30	30	30	30	20	20	NE
S. AFRICA	15	10	10	10	10	10	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	15	15	15	E
CARIBBEAN S. AMERICA	15	10	10	10	10	10	10	10	15	15	15	15	15	20	20	20	20	20	20	20	20	20	15	15	SE
ANTARCTICA	15	15	15	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	15	15	S
NEW ZEALAND	10	10	10	10	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	15	15	15	10	10	SW
OCEANIA AUSTRALIA	10	10	15	15	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	15	15	10	10	10	W
JAPAN	20	20	20	20	30	30	30	30	30	40*	40*	30	30	30	30	30	30	30	20	20	20	20	20	20	NW

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

```
  :
652 H2=-176+(1490/M2)
653 REM CALC T.O. ANGLE
654 Q5=ATN(H2/Q4)
655 Q6=Q5*R1
  :
695 PRINT "GREAT CIRCLE DISTANCE, XMTR TO RCVR=",Q1
696 PRINT "GREAT CIRCLE BEARING=",F,"OR",F1
698 PRINT "TAKE OFF ANGLE=",Q6
  :
1071 Q1=G1*R1*111.12
1072 J1=0
1073 IF Q1<4000 GO TO 1080
1074 J1=1
1075 Q2=Q1
1076 FOR J1=1 TO 6
1077 Q2=Q2-4000
1078 IF Q2<4000 GO TO 1080
1079 NEXT J1
1080 Q3=Q1/(J1+1)
1082 Q4=Q3/2
1085 K6=1.59*G1
  :
1115 F3=100
  :
1271 REM CALCULATE BEARING, XMTR TO RCVR, F
1272 E=(SIN(L2)-(SIN(L1)*COS(G1)))/(COS(L1)*SIN(G1))
1274 F=ACS(E)*R1
1276 F1=360-F
  :
1772 F2=G2/M9
1774 IF F2>F3 THEN 1779
1776 F3=F2
  :
1782 G3=G2
1784 M2=G3/F3
1786 IF M2>2.2 GO TO 1790
1788 M2=2.2
  :
```

fig. 2. Modified MINIMUF 3.5 program listing.

torial propagation will favor evening hours during periods of high solar flux and disturbed geomagnetic field conditions.

Thirty and forty meters are both day and night bands. Intermediate distances (1000 to 1500 miles or 1500 to 2200 km) in any direction represents daytime DX. Nighttime DX on these bands may be expected to offer greater distance paths than on 80 meters and, like 80, follow the darkness path across the sky. Reduced midday signal strengths and distances may occur on days of high solar flux values or periods of anomalous absorption, with 30-meter openings disappearing in the pre-dawn hours on the morning after a high solar flux value has occurred.

Eighty and one-sixty meters will exhibit short-skip propagation during the daylight hours and lengthen for DX at dusk. These bands follow darkness, opening to the east just before your sunset, swinging more to the south near midnight, and ending up in the Pacific areas during the hour or so before dawn. Except for daytime short-skip signal strengths, high solar flux values hardly affect these bands. On some days, however, the condition known as anomalous absorption will diminish day *and* night signal strengths. The 160-meter band opens later and ends earlier.

reference

1. Robert B. Rose, K6GKU, "MINIMUF: A Simplified MUF-Prediction Program for Microcomputers," *QST*, December, 1982, page 36.

ham radio

the high-tech repeater: designing and building an FM translator

Modular construction simplifies assembly, operation, and maintenance

In "Linear Translators," (*ham radio*, September, 1983, page 14) James Eagleson, WB6JNN, described narrowband techniques useful in the design of improved FM repeaters. This article extends that discussion to include the design of lossless feedback amplifiers with low SWR and low noise figure; theory, design and performance of the valid signal detector; techniques for identifying a translator; and site selection and sensitivity/transmit power considerations for putting up a co-channel repeater. — **Editor.**

The 146.34/94 repeater designed and built by the Sierra Amateur Radio Club in Ridgecrest, California, is not actually a repeater, but rather an FM translator.

In a generic sense, translators *are* repeaters, in that both perform similar functions. Both translators and repeaters receive signals on one frequency and retransmit them, with increased power, on a second frequency.

There are important differences, however. A repeater is a transmitter with its audio input connected to the audio output of a receiver. A translator, on the other hand, heterodynes the received signal to the intermediate frequency (IF), amplifies it, and then heterodynes it to the transmit frequency. In a translator, the signal never exists as audio.

The primary advantage of true repeaters is that they can be built from surplus or otherwise easily obtainable parts. Translators — often representing the "state-of-the-art" — must be custom-designed and custom-built from parts that are likely to be more costly and perhaps difficult to find. But because translators offer improved performance and excellent signal quality, we opted to design a translator rather than a repeater and accept both the increased challenge and expense.

Design and construction of the translator was a shared effort. Although John Piri, WD6CSV, and I were responsible for overall system design and construction, Chuck Swedblom, WA6EXV, took charge of design and construction of the identifier. Ron Skatvold, WB6VXI, designed and built the transmit power amplifier. George Kreager, KB6HC, was responsible for construction of the transmit and receive filters, and Elvy Hopkins, ND6Q, designed and supervised construction of the antennas, feedlines, and masts.

Many others assisted with general planning, frequency choice, site selection, and innumerable additional details. The project, from initial conception to installation and operation, took approximately 10 months.

translator design: overview

The translator was designed and constructed as seven modules, as shown in the block diagram of **fig. 1**. These modules include the down-converter, the IF section, the up-converter, the power amplifier, the control module, the valid signal detector, and the identifier. The down-converter receives the signals

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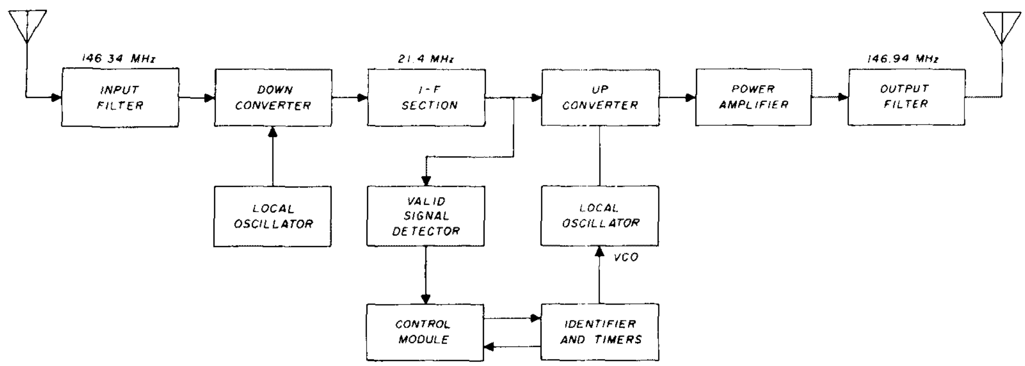

fig. 1. Block diagram of translator. No audio circuits are used in this 2-meter FM repeater.

passed by the input filter and heterodynes them to the 21.4 MHz IF. The IF section filters and amplifies these signals. A total of 10 poles of crystal filtering are included so that the only signals retransmitted are those originating in the translator's input passband. The IF section provides the major amount of amplification, with over 100 dB of gain. The up-converter heterodynes the IF signal to the transmit frequency. In addition, it has a crystal oscillator for inserting a substitute carrier during identification if there is none at the input. The power amplifier amplifies the 10-mW signal from the up-converter to the 10-watt level. The identifier controls the timing of the translator control and generates the identifier tones. The timers for identifying the translator, timing out the transmitter, and setting the code rate are included in the identifier. The identifier depends upon input from the valid signal detector. The valid signal detector samples the 21.4 MHz IF to measure the signal-to-noise ratio of the signal in the IF passband. When the signal-to-noise ratio is above a preset level, the transmitter is turned on and the identifier starts timing. The control module controls both the transmitter and the carrier insertion oscillator.

The sensitivity of the translator is very good — about −125 dBm (0.12 microvolt) at the receive filter. Limited transmit power and directional antennas control the coverage area of the translator.

operation

The site selection study was carefully performed, taking into consideration the club's desire to have a semi-local repeater, the lay of the land, the distribution of population centers, and the location of existing repeaters. Briefly, the area consists of a very large valley surrounded on all sides by mountains rising to peaks 5000 to 8000 feet high, with population centers small and sparsely located. To serve the community and the major highway nearby, and to keep the repeater truly local, the site selection committee recommended that the translator be located not on a mountain top, but 200 feet above the floor of the southern end of the Indian Wells Valley, using north-facing directional antennas. The site selection study showed that a balanced repeater — neither an alligator nor an elephant — with a transmit power of approximately 10 watts would cover the desired area.

The site selected allows the translator to serve Amateurs using handheld transceivers in the Ridgecrest area and those using proper mobile equipment transmitting 2 watts or more on the nearby highway for more than 50 miles. The only co-channel interference occurs with base stations operating in the overlap zone with a repeater to the north. No tone squelch is necessary because there are only one or two Amateurs in the overlap zone, and they generally use directional antennas.

The translator was designed to provide high-quality signal reproduction indistinguishable from simplex. (Indeed, if the user's transmitter is slightly off frequency, the translator output is off frequency a like amount.) On-the-air experience confirms that this goal was met; when the output is compared with the input, no discernible difference can be heard. In fact, to the first-time user, the identifier is the only clue that there is a repeater on the channel.

To keep the design simple the translator has a very short squelch tail, and no carrier tail at all. We were surprised to discover how many transient Amateurs were confused by the lack of the usual carrier tail. As an experiment we tried using the translator *with* a carrier tail. After a trail period of two months the club membership voted to remove the carrier tail; operation without it is so clean that the presence of the tail was actually a nuisance to local operators.

The translator provides reliable coverage for 50 miles

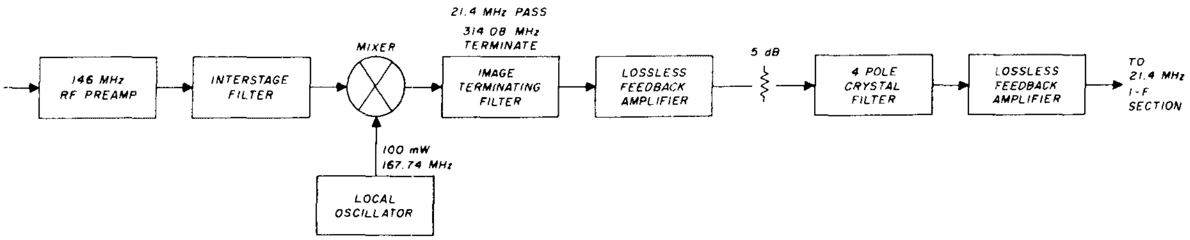

fig. 2. Down-converter block diagram. The down-converter is designed for wide dynamic range.

to the north. Because everything (except the backup-power batteries) is solid-state and designed for reliability, we have enjoyed excellent results. Periodic maintenance to service the nickel cadmium emergency power batteries and retune the crystal oscillators is scheduled once every two years. Only one failure, caused by the failure of a 2N2925 transistor in the control module, has occurred in seven years of service.

down-converter design

The down-converter, shown in **fig. 2**, consists of a crystal-controlled local oscillator, an RF preamplifier, an RF mixer, and the first stage of the IF amplifier with a 4-pole pair crystal filter. The IF chosen was 21.4 MHz because this was the highest frequency for which high-performance crystal filters were readily available. This module was designed for weak-signal sensitivity in the presence of strong interfering signals. Because the transmitter and receiver operate simultaneously, the potential for intermodulation distortion is ever-present; consequently, intermodulation performance was given high priority. *Failure to avoid intermodulation distortion causes repeaters to respond to signals at frequencies other than the designed input frequency.* We also wanted to design the translator for superior intermodulation performance in order to reduce the problems of keeping cavities tuned "just right."

For first-rate intermodulation performance it is essential to limit the gain before narrowband (crystal) filtering. For this reason, the preamp gain was limited to little more than 10 dB. In addition, a crystal filter was included as close as possible to the down-converter mixer as well as at the end of the IF chain. Lossless feedback was employed with the RF preamp to achieve low SWR and low noise figure simultaneously. The interstage bandpass filter has 1 dB of insertion loss. The interstage filter includes a notch at the image frequency and another notch at the most prominent spurious response frequency.

The RF mixer was also built from scratch. With a 100-mW LO the conversion loss was less than 5 dB, and the third-order intermod intercept point was +23 dBm. The image-termination filter following the mixer represents a 50-ohm match at 21.4 MHz and an open circuit at 146 MHz. It also includes a short circuit at the RF + LO frequency, which gives better intermodulation distortion than a 50-ohm match at that frequency.[1]

The crystal filter input impedance outside of its passband is significantly higher than the midband value. Reflecting a high impedance to an amplifier seriously reduces the third-order intercept performance of the amplifier. It is, therefore, desirable to transform to a low impedance the above-nominal impedance reflected to the amplifier for signals outside the crystal filter passband. The impedance inverting property of a properly designed Pi network is desirable for this purpose. Certain L networks will also provide the same above-nominal impedance transformation, but the Pi network has the advantage of easy tunability with both variable capacitors returning to ground. For these reasons, all crystal filter impedance matching networks were of this design.

When the down-converter was designed it did not include the lossless feedback amplifier and 5-dB attenuator. We found that the RF preamplifier and mixer could handle a weak signal in the presence of multiple high-level inputs. But the output impedance of the mixer would change from a 50-ohm match with input levels above −10 dBm (−20 dBm at the input to the RF preamplifer), thus seriously degrading the passband flatness of the crystal filter in the presence of a strong off-frequency signal — in particular, the repeater output. An interstage amplifier with low noise figure, low intermodulation distortion, and a constant output impedance was needed. To keep distortion down, a low-gain amplifier was needed, but that implies feedback to limit gain. To achieve low gain, low noise, and low distortion, a lossless feedback amplifier was the answer.[2] Unfortunately, such amplifiers pass the impedance seen at the input to the output with little isolation. The 5-dB attenuator (**fig. 2**) was included so that the crystal filter is presented with a constant impedance with all RF input levels. The amplifier was designed with 6-dB gain, a low value in the interest of improved intermod performance. The transis-

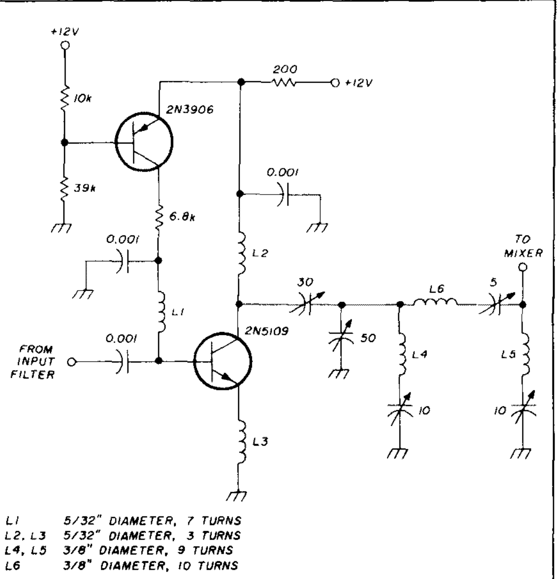

fig. 3. RF preamplifier schematic. Inductive reactance emitter degeneration is used to improve the preamp input SWR without significantly degrading the noise figure.

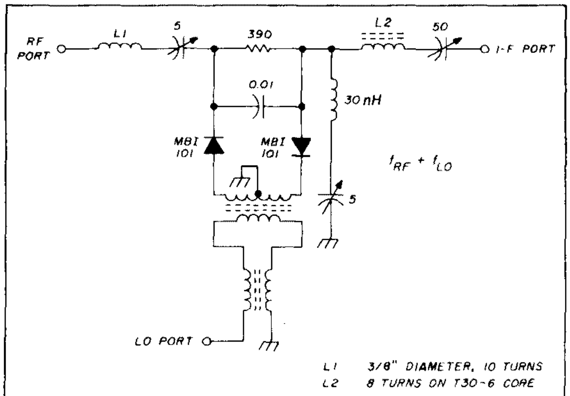

fig. 4. Single balanced mixer used in the down-converter and the up-converter. The transformers are wound on ferrite beads (μo = 64). The 30 nH and 5 pF variable capacitor comprise the f_{RF} + f_{LO} (314 MHz) termination impedance.

tor was biased for high third-order intermodulation intercept, and yielded better than a 3-dB noise figure. This design provides a good match for preserving the crystal filter passband flatness, and has low noise figure. Unfortunately, it does degrade the intermod distortion performance slightly.

Noise figure and gain calculations showed that the noise figure of the first stage following the crystal filter was important. The best noise figure attainable with conventional amplifiers was approximately 5 or 6 dB, which was insufficient. The problem was that the crystal filter must have an impedance match to have a flat passband response. However, having a match with a conventional amplifier guarantees a poor noise figure.[3] Again the solution was a lossless feedback amplifier for the IF amplifier following the crystal filter. By favorably biasing the amplifier for improved noise figure, a 2-dB noise figure was attained. A conventional amplifier was cascaded with the lossless feedback amplifier for a net gain of 30 dB in the down-converter.

The local oscillator (LO) for the down-converter (and the up-converter) uses double-tuned interstage filtering to keep all spurious outputs in the LO output below 60 dBc.[4] The local oscillator puts out 100 mW of LO power to the mixer.

Figure 3 is the schematic of the RF preamplifier. The 2N3906 was included to control the collector current through the 2N5109 independently of temperature. The 146-MHz preamp has 11 dB of gain. It achieves a 3-dB noise figure with an input SWR of approximately 2:1. The SWR of the preamp must be low so that the response of the RF input filter, which precedes the preamp, will not be distorted. Low SWR and low noise figure were achieved with lossless feedback. The lossless feedback was obtained with inductive reactance in the emitter of the RF transistor,[5] which in turn determined the SWR. An SWR of nearly 1:1 is possible, but at the expense of the noise figure. The value of inductance chosen was a compromise between low noise and low SWR.

Figure 4 is the schematic of the RF mixer. It was designed to convert an RF signal to the IF. To keep the intermodulation distortion down, 100 mW of LO power was applied to the mixer. The 390-ohm resistor develops a back bias voltage from the LO current flowing through the diodes. This bias voltage increases the reverse voltage for the off state of the diodes, which allows a greater peak-to-peak RF input level for the same distortion level. A single balanced mixer, it requires a diplexer to separate the RF from the IF, which share a common port. Generally a diplexer is composed of high-pass and low-pass filter sections sharing the common port. However, it was considerably easier to design and tune a diplexer made from series-tuned filter sections. This is acceptable in a frequency converter for use in a repeater, where RF bandwidth is not a consideration.

IF module

The IF (21.4 MHz) was chosen to take advantage of readily available high-quality crystal filters. Although no problem has been observed, a higher frequency

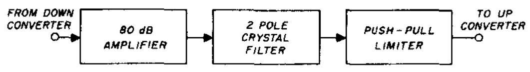

fig. 5. IF module block diagram. This module is physically located in two separate enclosures. An additional 25 dB of IF gain is included in the down-converter enclosure.

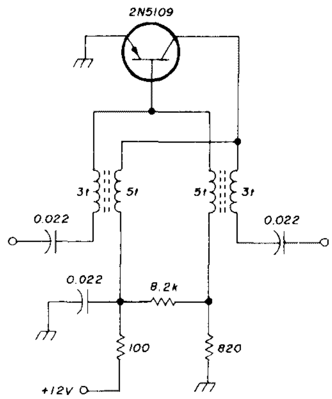

fig. 6. 6 dB lossless feedback IF amplifier. For higher gain reduce 3 turn windings to 1 or 2 turns. For lower noise figure increase 100 ohm resistor to 330 ohms. The 0.022 μF capacitors are very small monolithic ceramics.

would have made image rejection and spurious response suppression more effective. The IF section block diagram is shown in **fig. 5**. More than 100 dB of IF gain was used, resulting in 20 dB of excess gain before limiting. The excess gain was desired to ensure that the limiter was hard limited under all conditions. It was not possible to put that much gain in one box without the risk of oscillations from feedback. To prevent that from happening, the gain stages were distributed among three separate enclosures. The use of double-shielded interconnecting coax cables also minimized coupling.

Figure 6 is the schematic of the lossless feedback IF amplifier used on each side of the crystal filter. The two amplifiers differ only in the turns ratios of the transformers and the bias currents of the transistors. To keep the circuit physically small, the transformer cores were 0.100-inch diameter ferrite toroids with a permeability of 125. Considerable care must be used in the layout of this amplifier so that stray reactances do not degrade performance. This amplifier design is useful to about 50 MHz.

The translator includes a limiter in the IF to maintain constant power level to the up-converter mixer. A 2-pole-pair crystal filter precedes the limiter to prevent off-frequency IF amplifier noise from suppressing a weak signal in the limiter. The effectiveness of a limiter is degraded if limiting is not symmetrical; that is, if limiting of the positive and negative swings of the IF waveform are not equal. To achieve symmetrical limiting, a push-pull limiter was employed. By ensuring that the limiter is limited even on noise, there is no variation in translator performance with signal strength, supply voltage, or ambient temperature.

We decided not to include SSB signal-handling capability for this repeater. The only changes necessary for passing SSB would be to use AGC instead of limiting, and to use a linear transmit amplifier. A limiter was chosen to provide constant signal amplitude to the up-converter with minimum circuit complexity. If AGC were to be used instead, the amplitude components of a weak signal would have required a linear power amplifier to keep intermodulation products from splattering across adjacent channels.

up-converter

Figure 7 is the block diagram of the up-converter. A 4-pole crystal filter was included just before the IF signal was up-converted to the transmit frequency. This guarantees that the only signals transmitted are within the transmit channel bandwidth.The up-converter heterodynes the 21.4 MHz signal to 146.94 MHz.

Because there are no audio circuits in the translator, the repeater cannot be identified in the usual manner of inserting audio tones into the modulator. In this translator the identification modulation is accomplished by frequency modulating the crystal-controlled local oscillator in the up-converter with the identifier tones. This conveniently adds the identifier tones to the user's signal. If no carrier is present during the identification time, a substitute carrier is inserted from a crystal oscillator.

The up-converter mixer is identical to the mixer in the down-converter. It is very important to have a low distortion mixer in the up-converter so that spurious outputs are below the FCC limits. The IF amplitude was adjusted, with the limiter bias current, until the spurs were sufficiently below the FCC limits. This resulted in approximately 1-mW output from the mixer. Several stages of 146-MHz amplification were required to raise the 1 mW to the final 10-watt level. Most of that gain was provided by the power amplifier module. The power amplifier was designed as class C for high efficiency.

When there is no input to the repeater, the output from the crystal filter in the up-converter is full-power noise. If the control module calls for the translator to

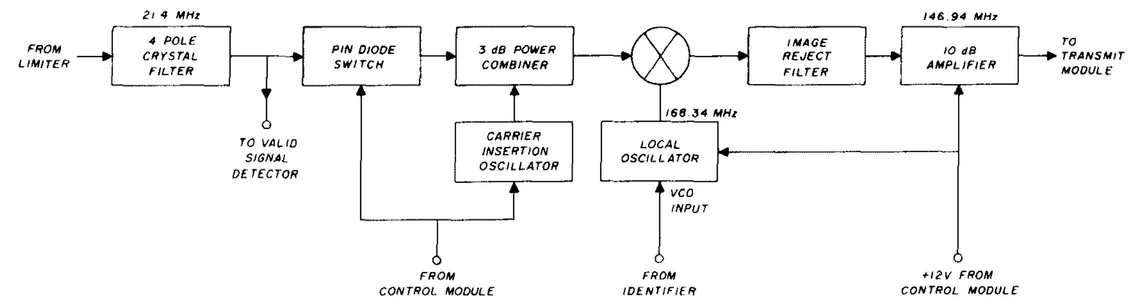

fig. 7. Up-converter block diagram. A high-level mixer is necessary to keep spurs low. The PIN diode blocks noise out of the IF if no carrier is present.

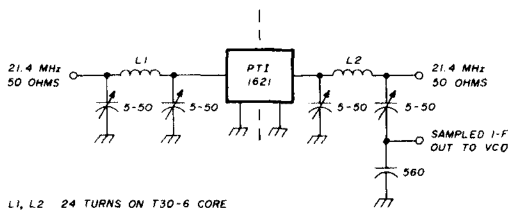

fig. 8. Piezo Tech 4 pole-pair crystal filter and matching networks. Found in up-converter. The 560 pF capacitor is omitted in the down-converter version. Careful tuning resulted in better than ±0.1 dB passband flatness.

be identified when there is no input signal, the carrier insertion oscillator is turned on. Summing the full-power noise and the carrier insertion oscillator signal would result in equal carrier power and noise power, which would give a 3-dB signal-to-noise ratio for the identifier. A PIN diode was used to block the noise coming out of the IF when the carrier insertion oscillator is on. This diode attenuates the IF noise sufficiently to yield about a 30-dB signal-to-noise ratio when the carrier is inserted locally.

Figure 8 shows the up-converter crystal filter matching network. A small amount of the IF signal is sampled at the output of the crystal filter for use by the valid signal detector circuit. Sampling with a low-impedance tap minimized the effects of the high SWR of the PIN diode when it conducts.

identifier/control module

The control module controls the repeater operation. In addition to its obvious tasks, it controls the PIN diode switch and the carrier insertion oscillator to insert a substitute carrier when needed for identification. The control module receives its inputs from the valid signal detector and the time-out timer. The valid signal detector, described in the following section, determines the signal-to-noise ratio of received signals. When the signal-to-noise is above a preset level, the transmitter is turned on and the time-out timer is started. The timer is reset on loss of carrier.

The identifier controls the timing of the repeater and generates the identifier tones. The identifier uses the valid signal detector output for control. The timers for identifying the repeater, timing out the transmitter, and setting the code rate, were included in the identifier. Control for the timers comes from the control module.

valid signal detector

The valid signal detector samples the 21.4-MHz IF signal to measure the signal-to-noise ratio in the IF passband. The signal-to-noise ratio is measured by evaluating the amplitude variations in the IF signal. The IF signal consists of frequency modulated (FM) signals superimposed on amplitude modulated (AM) signals. The FM signals are the desired signals and the AM signals are undesired, usually noise. It can be easily appreciated that a clean FM signal will have no AM components in it, and that noise is composed of AM and FM components. What is not so evident is the strong AM components of noise after the signal is hard limited and bandpass filtered. A hard limiter converts any signal, even one with significant amplitude variations such as noise, into a constant power, wideband signal with no amplitude variations. After bandpass filtering to the original frequency and bandwidth, the signal will have constant *power*, but not necessarily constant amplitude.[6] A noise-free FM signal will be a constant amplitude (and constant power) frequency-modulated sine wave, but noise will be band limited noise with fluctuating amplitude with the same

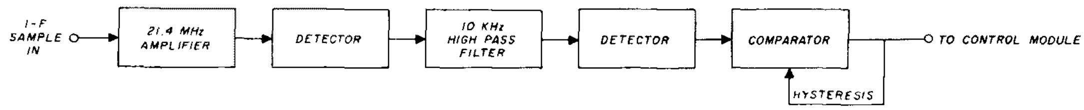

fig. 9. Valid signal detector block diagram. This circuit measures the signal-to-noise ratio in the IF.

average power level. After bandpass filtering then, hard-limited noise will once again have amplitude variations similar to the original noise. It is this characteristic that the operation of the valid signal detector depends upon.

Figure 9 is the valid signal detector block diagram. The IF signal is sampled after the post-limiter crystal filter. The valid signal detector amplifies the IF signal and provides envelope detection. The envelope detector furnishes a voltage proportional to the instantaneous IF signal amplitude. As a result of the limiter, the IF power applied to the detector is constant. If no carrier is present, full power noise is applied to the detector, which produces a fluctuating DC voltage. In this case, the frequency components of the detector output extend from DC to 15 kHz, the IF bandwidth. When a strong carrier is introduced, the limiter suppresses the noise so that the output of the detector becomes a noise-free DC voltage. It can be seen therefore, that the high-frequency components of the detector output correspond to the sidebands of a carrier, either AM sidebands or noise. In theory, and in practice, any incidental AM on an FM signal is limited to roughly 3-kHz bandwidth. Therefore, the signal-to-noise ratio of a received signal can be determined by measuring the amplitude of the 10- to 15-kHz components of the envelope-detected signal.

The components resulting from the noise level are extracted by high-pass filtering of the detected signal. The detector following the high-pass filter then provides a DC voltage level inversely related to the signal-to-noise ratio. Actually, the DC level is equal to a constant minus the received signal strength, but for practical purposes it provides a very good approximation to signal-to-noise ratio. The threshold for the comparator in the valid signal detector has significant amount of hysteresis. Hysteresis was included so that any signal that brings up the repeater is workable, but a fading signal will not be cut off until the bitter end.

The valid signal detector has provided outstanding service. It is so effective that even weak signals reliably bring up the translator, but noisy, uncopyable signals are virtually never heard. The only problem that occurs is with signals with severe multipath distortion. Multipath distortion can create a poor signal-to-noise ratio during modulation, even though the signal has acceptable signal-to-noise ratio when not modulated. This occurs when destructive interference causes notches in the signal amplitude at several frequencies in the modulation bandwidth. The result is a drop in the signal strength during modulation, and the reduced strength of the modulated signal does not have sufficient signal-to-noise ratio to keep the translator on. When this happens, the person transmitting brings up the translator whenever he or she pauses, but gets cut off as soon as he or she speaks. Such signals are unintelligible anyway; it just sounds like the valid signal detector is unnecessarily cutting off a copyable signal.

input/output filters, antennas

The RF input and output filters were constructed from a design put forth by Tilton.[7] The filters provide a 50 dB notch at the rejection frequency with a 1 dB insertion loss. This is identical to the performance predicted in the reference, even though silver plating was not applied as in the original filters. To enhance the output-to-input isolation for the repeater, separate transmit and receive antennas, spaced about 70 feet apart, were used. The antennas are surplus commercial five-element Yagi antennas. To retune to the Amateur band, we set up a ground effect antenna range[8] and found we could achieve the same gain as the original five-element antennas with only four elements, using the original boom. The side lobes with four elements were slightly stronger than with five elements, but still met our requirements.

squelch/carrier tail considerations

A few comments on the operational characteristics of this repeater are included here for the benefit of any who might seek to advance FM translator design. When the translator was designed we understood that pure noise out of the IF would cause intermodulation products in the class C power amplifier. For that reason the valid signal detector squelch response time was made short — around 50 ms. When someone listens to a weak signal on a frequency up to 1 MHz away while there is a conversation going on over the translator, the noise burst at the end of each transmission through the translator blanks out the weak sta-

tion for the duration of the squelch tail. It turns out that this phenomenon is not particularly noticable: it has not been noticed, even in locale in which almost all 2-meter activity is weak-signal work.

It appears that the main annoyance is the translator's emission of a full-bandwidth (15 kHz) noise burst at the end of each transmission. The typical squelch circuit in the user's receiver responds to this noise burst in the same way it would if there were no carrier on channel and squelches the receiver audio. That is then followed by a clean carrier tail that unsquelches the radio. After the carrier tail times out, the user's radio again squelches. The result at the user's receiver is two squelch bursts for one end-of-transmission. (This phenomenon does not occur with a conventional repeater because its audio stages limit the bandwidth of the noise burst so that the typical squelch circuit will not respond.) Solving this problem with translators is not easy; it will occur with class C and linear power amplifiers alike. One temporary solution is to make an inordinately long carrier tail so that the second burst is likely to be taken up by another user; any suggestions from readers for providing a clean carrier tail would be appreciated.

conclusion

This article has outlined the objectives and results of a project for the design and construction of a translator/repeater. Designed and constructed entirely "from scratch," the repeater has met all performance goals and has operated with minimal service for 7 years. Any comments or suggestions from readers will be appreciated. If you wish to receive a response, send an SASE to the author at the address at the beginning of this article.

acknowledgement

I would like to thank Bill Maraffio, N6PR, for reviewing this article manuscript form and for making many helpful suggestions.

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ham radio

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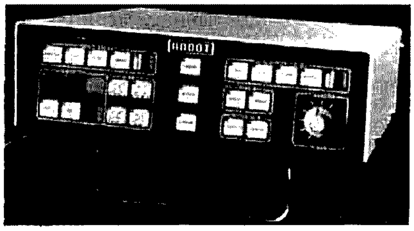
NEW products

hi-res color SSTV converter

A new high-resolution color SSTV converter has been added to Robot's line of Amateur Radio products. Designated the Model 1200C, it is capable of transmitting color video images said to rival broadcast television in picture quality. The Model 1200C has three selectable 6-bit memory planes that combine to form 262,244 color combinations in a 256 × 240 line full screen display.

Eight different black and white and color transmission formats are available with automatic selection on receive. Up to six separate pictures may be stored in memory. The unit accepts color or black and white composite video from standard TV cameras and has RGB, composite or RF modulated video output.



One distinctive feature of the Model 1200C is the 8-bit parallel I/O ports for computer interface. This allows total access to each individual pixel by a host computer for image processing, transformation, storage and recall, and graphics. This port also allows the connection to a printer for black and white or color hard copy picture printing.

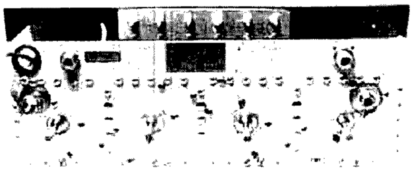
The Model 1200C features touch sensitive front panel switches for full station control and several automatic functions for unmatched ease of use. Fine tuning, speed switching and color or black and white detection are automatically accomplished without operator intervention.

For further information, contact Robot Research, Inc., 7591 Convoy Court, San Diego, California 92111.

Circle #175 on Reader Service Card.

bandpass duplexer

Sinclair Radio Laboratories' new P-4440E combines the low loss of a Res-Lok aperture-coupled filter on transmit with the high selectivity of a combiner filter on receive, making it an appropriate choice for single antenna operation of trunking or cellular base stations. The Res-Lok four cavity bandpass section has typically 0.5 dB insertion loss and provides nearly 50 dB of noise suppression, which, when added to the typically 35-40 dB noise suppression provided by most cavity ferrite transmitter combiners, provides nearly 90 dB noise suppression overall. The six-pole comb-line filter on the receive side provides



over 85-dB carrier suppression, a figure which is more than adequate for most system applications. In addition, both the transmit and receive bandpass windows are a full 15 MHz wide, a fact which allows sparing of this component on a multi-site basis without the need to have facilities for retuning.

For further information, contact Sinclair Radio Laboratories Inc., 675 Ensminger Road, Tonawanda, New York 14150.

Circle #312 on Reader Service Card.

remote coax switch

A remote coax switch for convenient switching of up to four antennas is now available from Heath. It mounts easily on a tower or mast and consists of a remote RF switching unit and an indoor control unit. The two units are connected by a single coaxial cable which handles both RF and control signals. The remote unit is rain tight and mounts with a single clamp. The control unit contains the power supply and provides switching signals to the remote unit. The HD-1481 switch handles 2000 watts PEP with a VSWR of 1.15:1 or less below 30 MHz.

For complete details and/or to receive a free copy of the latest Heathkit catalog, contact Heath Company, Benton Harbor, Michigan 49022.

Circle #308 on Reader Service Card.

"Decode-A-Pad"

The Engineering Consulting touch-tone to RS-232-C interface for home computers allows reception of all 16 DTMF touchtones as fast as they can be transmitted. The computer does all the work at 300 baud; each digit is displayed as it is transmitted. With the Decode-A-Pad, you can receive coded strings, decode any number of digits, and program as many multi-digit codes as you want — all in BASIC. Sample programs to get you started are included in the price.

Now you can use your handheld radio to control your computer, which can then be used to control your remote base or turn on and off relays, for example.

NEW products

The Model DAP1 is priced at $89.95, which includes domestic shipping.

For details, contact Engineering Consulting, 583 Candlewood Street, Brea, California 92621.

Circle #107 on Reader Service Card.

vented actuator boot

The new QIK-PRO Boot from Paullin Industries offers improved actuator protection against condensation with new top and bottom "flow-thru" vents to relieve moisture. These boots, custom-designed for satellite actuators, are made of neoprene rubber with self-retrieving folds for longer life and improved ice removal. The extra-long life is the result of a heat-curing process that provides the highest infra-red, ozone, and ultraviolet test ratings.

The QIK-PRO Actuator Boot has been redesigned to fit 1-1/8 x 2-1/4-inch actuators without adapters. Two ties are provided to seal the boot tighter, protecting the actuator equipment against moisture, dust, ice and rain.

For more information, contact Paullin Industries, 1446 State Route 60, Ashland, Ohio 44805.

Circle #307 on Reader Service Card.

new Hamtronics® catalog

The 1985 mail order catalog from Hamtronics features 40 pages of items of interest to the VHF/UHF/OSCAR enthusiast. Both new products – including a simplex autopatch kit, a repeater COR with courtesy beep, GaAs FET receiver preamps, active antennas for scanners, and repeater PA kits – and the firm's already popular lines of FM and AM receivers and 800 MHz scanner converters are described.

For a free copy, contact Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535. (For overseas mailing, please send $2 or 4 IRCs.)

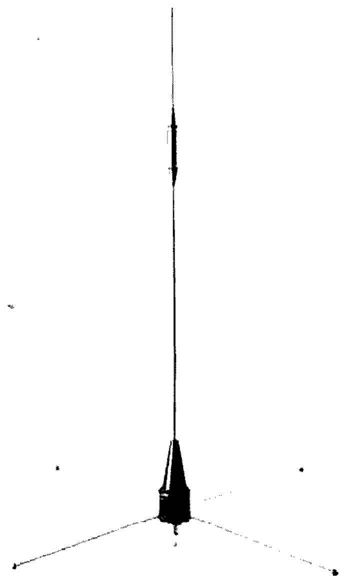
Circle #154 on Reader Service Card.

Larsen's FB2-450

Larsen's new FB2-450 antenna features an exclusive Kulrod T™ teflon-coated finish for corrosion resistance and improved performance. The lightweight, fixed-base antenna is said to offer a reliable, economical alternative to larger base station antennas for a variety of base station applications.

The 5/8 over 1/2 wave collinear whip with four integral ground plane radials delivers 4.5 dB

gain, compared to a 1/4-wave antenna on a suitable ground plane. The FB2-450 operates in the 450-470 MHz frequency range and ensures low loss with "N" type hardware. It comes complete



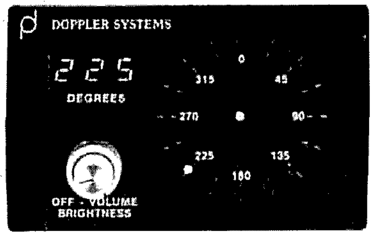
with bracket and hose clamps for mounting on a 1-1/2 to 2-1/2-inch diameter pole.

For more information, contact Larsen Antennas, P.O. Box 1799, Vancouver, Washington 98668.

Circle #306 on Reader Service Card.

direction finding accessory for VHF/UHF receivers

Doppler Systems' latest line of radio direction finding units operates with any narrowband FM receiver in the 27 to 500 MHz range to provide fast location of interfering signals. No receiver modifications are required — the direction finder connects to the receiver's antenna and external speaker jacks.



Four modes are available, providing a range of optional display and remote output features: a 16 LED compass rose; a 3-digit bearing in degrees; an RS232C output for computer analysis, triangulation, data logging, etc.; and

short circuits

microstrip program

In K8UR's "Microstrip Impedance Program," (December, 1984, pages 84-86) corrections should be made to lines 230 and 280 of **fig 1. Line 230,** ER [– .0724 should read ER I – .0724. In **line 280,**] – .5) should read I – .5). **Line 80** in the HP67 program listing (**fig. 2B**) should read:

hy^x	35 63

trap antenna

In W4MB-s article "design your own trap antenna" (October, 1984, page 37) the formula for Z_n should be corrected to read as follows:

$$Z_n = 60\ (\log_e \lambda_n/D_n - 1)$$

Also, do not confuse LN with (L(N). They are different. LN = natural logarithm to the base e.

NEW products

a synthesized voice output. The speech synthesizer is designed for mobile use, as it eliminates the need for the driver to watch the display.

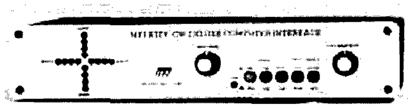
The system operates by continuously summing the outputs of four antennas, simulating the motion of a single rotating antenna. As the simulated antenna moves toward the RF source, an increase in the apparent signal frequency occurs, and as the antenna moves away from the source, this frequency decreases. This up-down (Doppler) frequency shift is detected by the FM receiver and is present as a 300 Hz tone on the audio output. The phase of the tone is measured and used to compute the bearing without affecting the normal operation of the receiver.

For more information contact Doppler Systems, 5540 E. Charter Oak, Scottsdale, Arizona 85254.

Circle #305 on Reader Service Card.

RTTY/CW computer interface

A new RTTY/CW deluxe computer interface, featuring variable tuning for all shifts and built-in RS-232 compatibility FM-AM modes of operation, is available from MFJ Enterprises, Inc.



The MFJ-1229 features a 16-LED crosshair mark and space tuning array that simulates a scope ellipse for easy, accurate tuning even under poor signal-to-noise conditions. It also operates in both FM and AM modes, using FM for general use, off-shift copy, drifting signals and moderate signal and QRM levels, and AM for weak-signal conditions or when there are strong stations nearby. The MFJ-1229 transmits on both 170 and 850 Hz with variable shift tuning as well as push-button 170 Hz for added convenience and versatility.

The 1229 can be used with most home computers and with a large variety of software.

Additional features include AFSK and FSK keying, front panel sensitivity control, a normal/reverse switch that eliminates retuning while checking for inverted RTTY, mark and space outputs for true scope tuning, and a Kantronics-compatible socket.

All inputs are buffered and can be inverted using an internal DIP switch. External trim pots are

accessible from the rear, allowing adjustment of the audio input levels. The MFJ-1229 uses 18 VDC or 110 VAC with an optional AC adaptor. The interface measures 12-1/2 x 2-1/2 x 6 inches and is housed in a black aluminum cabinet with a brushed aluminum front.

For more information on the MFJ-1229 Deluxe RTTY/CW Computer Interface, contact MFJ Enterprises, Inc., 921 Louisville Road, Starkville, Mississippi 39759.

Circle #173 on Reader Service Card.

Great Circle slide

Xantek has announced the availability of a Great Circle Slide for its DX EDGE®. Used with the DX EDGE, this slide lets stations determine beam headings (great circle bearings) to any location in the world with enough accuracy for almost any purpose. It also shows the beam heading to use for pointing an antenna along the Gray Line.

Slides are available for latitudes of 60, 50, 40, 30, 20, and 10 degrees (all north or south), and 0 degrees (the equator). (Order the slide for the latitude nearest your station.)

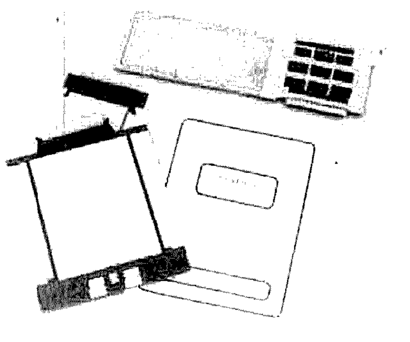
Each slide shows 16 true great circles spaced at intervals of 22-1/2 degrees. Both the short path and long path are shown. The same size as the DX EDGE map, the slide is made of thin transparent vinyl so that it fits over the map and under the monthly slide. The price of the Great Circle Slide is $3.00 when purchased together with the DX EDGE and $5.00 when purchased separately. Prices include shipping and handling.

For further information, contact Xantec, Inc., P.O. Box 834, Madison Square Station, New York, NY 10159.

Circle #304 on Reader Service Card.

interface development system

The 'eZ SYSTEM' is a low cost, simple and practical hardware development system that provides quick access to a personal computer's bus expansion slot for rapid circuit development. The system features three major components: the 'eZ BOARD,' a solderless prototyping board that connects to the expansion slot of the micro-computer through an integral 18-inch flat cable, allowing the user to work freely without interfering with their system unit; the 'eZ CARD,' a prototyping board that features a fully buffered address, data and control bus, and the 'eZ BOOK,' a helpful technical guide to the computer system and contains several practical, useful circuits for projects such as an A/D conversion, parallel port, and joystick interface.



The eZ System is available for the IBM-PC and XT, for Apple, and Commodore computers, or other computers housing the same bus arrangement.

For further information, contact Sabadia Export Corporation, P.O. Box 1132, Yorba Linda, California 92686.

Circle #302 on Reader Service Card.

do-it-yourself autopatch

KIE Enterprises has developed a simplex telephone autopatch system called "LETUS-PATCH" for the VIC-20 or Commodore-64 computers. The system (not including the cost of the computer or the transceiver) can be built for less than $50.00.

Features of the autopatch include multiple-user "up codes" with individual long-distance privilege designation, general admission "up codes," busy phone and ringing phone indication (on or off). Also included are answer or interrupt busy phone "up codes," automatic ID'er, remote enable/disable capabilities, and voice delay switching of the phone conversation for those with slow response transceivers.

For those whose telephone service no longer requires dialing a "1" before the area code, a prefix/area code table per "up code" can be added. All phone numbers used are logged onto cassette tape by "up code," date, and time to comply with FCC requirements and to allow multiple user toll charge distribution when cross-referenced with your phone bill.

The program, written in BASIC, requires approximately 3K of memory. Extensive prefix or area code tables may require memory expansion for the VIC-20.

The "do-it-yourself" package consists of circuit diagrams, circuit descriptions, parts lists, program listing, program narrative, and a cassette tape with a starter program ready to load. The price is $20.00.

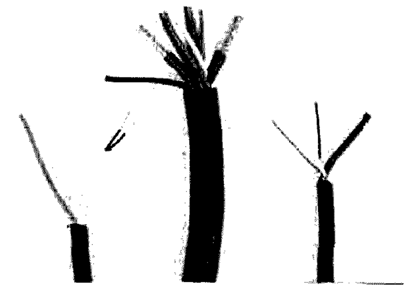
For more information, contact KIE Enterprises, P.O. Box 72, Running Springs, California 92382.

Circle #301 on Reader Service Card.

actuator cable

Nemal Electronics International has introduced a new line of direct burial actuator cable for satellite earth station and communications applications. Each type provides the proper cabling for both motor power and sensor/control in a single polyethylene jacket suitable for direct

NEW products



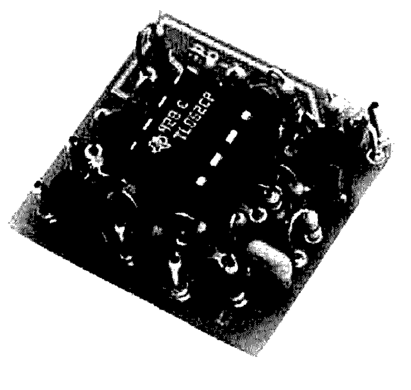
burial. Nemal ST-1 consists of 2 conductors of 16 gauge and 2 conductors of 22 gauge with foil shield and drain wire; ST-2 contains 2 conductors of 12 gauge and 3 conductors of 22 gauge with foil shield and drain wire. Nemal also offers a line of five types of satellite control cables which contain motor, sensor, polarotor, and coaxial signal lines.

For additional information, contact Nemal Electronics International, Inc., 12240 N. E. 14th Avenue, North Miami, Florida 33161.

Circle #166 on Reader Service Card.

reverse burst accessory

Communications Specialists has introduced the RB-1 *reverse burst accessory*. The RB-1 eliminates the long squelch tail heard with some reed type and other sub-tone decoders. Used in conjunction with decoders that offer squelch tail elimination, the RB-1 will delay the transmitter turn-off time and reverse the phase of the en-



coded tone, immediately stopping the decoder and eliminating the squelch tail. The RB-1 is available from stock and sells for $14.95.

For more information, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665-4296.

Circle #120 on Reader Service Card.

8-pole crystal filters

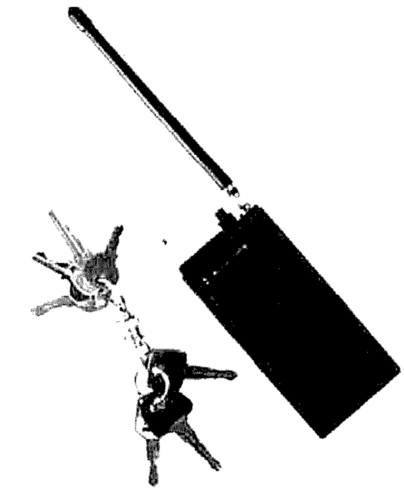
International Radio Inc. has announced its line of 8-pole crystal filters designed to improve the selectivity in Kenwood and ICOM products. Designed with low insertion loss and ripple, the filters offer excellent selectivity and shape factor. Filers are available for both SSB and CW operation, and depending on the radio, can be either 1st or 2nd IF or cascaded.

For more information, contact International Radio Inc., 1532 S.E. Village Green Drive, Port St. Lucie, Florida 33452.

Circle #109 on Reader Service Card.

VHF FM monitor receiver

Ace Communications, Inc. has introduced a new VHF FM monitor receiver, model AR-33. The AR-33 is a featherweight microprocessor controlled VHF FM portable receiver covering the 140 to 170 MHz band in 5 kHz steps. Frequencies are selected by a thumbwheel switch and slide switch for 5 kHz increments. The receiver employs CMOS microprocessor technology to offer a variety of features at an economical price, as well as small size and light weight. The actual size of the receiver is approximately 5-1/4 × 2-1/2 × 1-1/8 inches (130 × 63 × 26 mm) and weighs less than 200 grams.



For further information, contact Ace Communications, Inc., 22511 Aspan Street, Lake Forest, California 92630.

Circle #164 on Reader Service Card.

NEW products

voice operated squelch

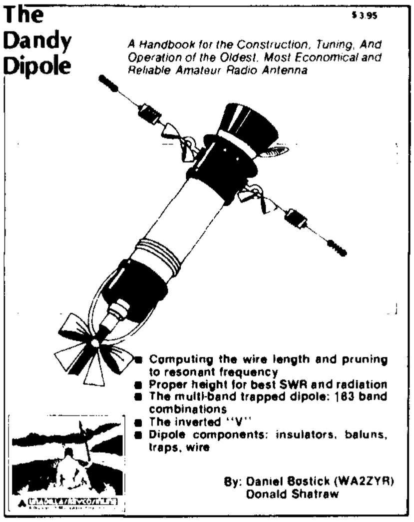
CMC Communications offers a Voice Operated Squelch on a small circuit board for mounting inside most SSB transceivers and receivers. The VOS requires that different and select components of the voice spectrum to be present at the same time to operate. It ignores heterodynes and noise, regardless of level, yet it is extremely sensitive to weak signals when the human voice is present. All adjustments are made at the factory and a remote on/off switch is provided. Simple connections are made to the AF Gain Control and 9 or 12 VDC.

For further information, contact CMC Communications, Inc., 5479 Jetport Industrial Blvd., Tampa, Florida 33614.

Circle #125 on Reader Service Card.

dipole handbook

The Dandy Dipole from Microwave Filter is a new 24-page handbook for constructing over 180 variations of the oldest, most reliable, and simplest Amateur Radio antenna known. It shows where and how to place it, how to quickly design a multiband dipole — using traps — without guessing at the wire lengths.



The contents include computing the wire length and pruning to resonant frequency; proper height for best SWR and radiation; the multiband trapped dipole, 183 band combinations (complete wiring tables); inverted "V" and dipole components as insulators, baluns, traps and wire.

The book was written by Daniel Bostick, WA2ZYR, and Donald Shatraw, Microwave Filter Company technical consultants.

The cost is $3.95 plus $1.00 for shipping. For more information, contact Microwave Filter Company, Inc., 6743 Kinne Street, East Syracuse, New York 13057.

Circle #148 on Reader Service Card.

book REVIEW

The Hidden Signals on Satellite TV by Tom Harrington, W8OMV, and Bob Cooper, VP5D

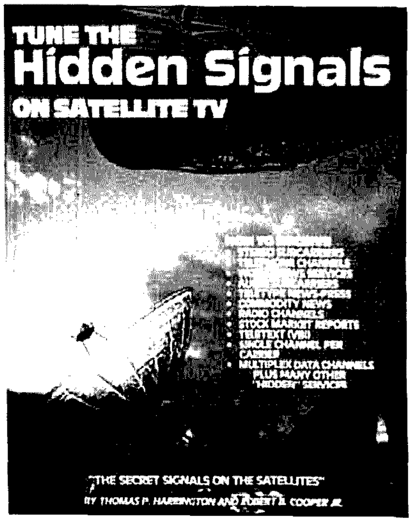
Owners of TVRO systems may be unaware of the multitude of signals carried by geostationary satellites. Besides the video services, these signals include audio and teletype news services, high speed data systems, teletext, and stock services, to name just a few.

Access to these signals is actually very easy. It can be accomplished simply by tuning the audio subcarrier control on many of the TVRO tuners currently being sold. With the addition of an FM stereo tuner, high quality stereo services can also be received and enjoyed by TVRO owners.

This book starts off with an introductory chapter that gives the reader a brief history of satellite communications, describes multiplexed audio and data signals, and lists the various types of transmissions. A spectrum overview from 0-30,000 MHz and a description of where the "hidden signals" are also provided.

Chapter Two delves into the technical "nitty gritty" of audio subcarriers, how they get there and how they can be found. In addition to a technical description, a review of several different pieces of commercial gear currently available to decode this subcarrier information is provided. A program listing by service and satellite is also included. Charts, graphs, and oscilloscope photographs are all provided to help you set your station.

Chapter Three discusses telephone (SSB/FDM) systems, their use, and your responsibilities in receiving these services. There are currently two types of voice and data signal channels on satellites these days. The most common uses SSB run through a frequency division multiplexer. This spectrum-efficient system allows many different signals to be combined together into a single signal for transmission to a satellite.



The second method of access is via SCPC (single channel per carrier). SPSC is not as spectrum effective as the FDM systems, but is less expensive to install. The equipment required is much less complex and needs relatively low power. Information is given on how to hook up SSB receivers to TVRO receivers to decode this interesting information.

Chapter Four is a rather complete and inclusive section on SCPC satellite systems. Equipment in use is fully covered and examples are given of some of those services using SCPC transmissions. Frequency allocations are listed by satellite, service, companding, bandwidth and pre-emphasis to help the listener tune in these signals.

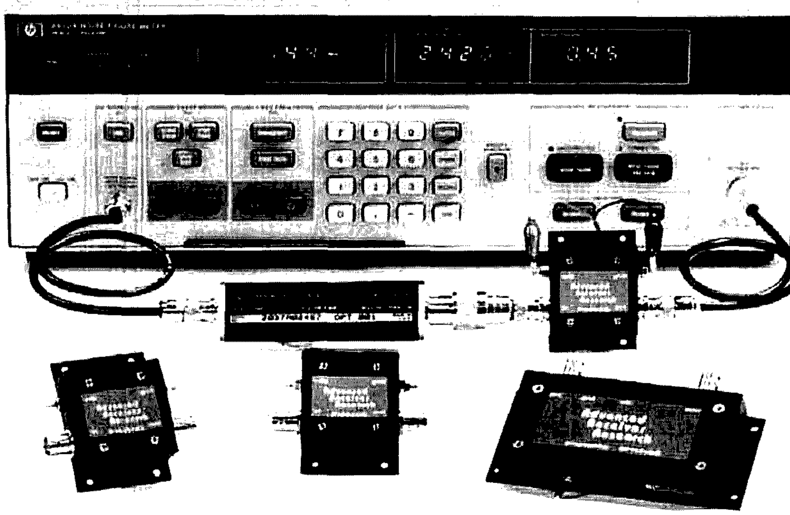
Chapters Five and Six cover satellite networks and basic teletext, providing plenty of history as well as technical details. Chapter Seven deals with a number of other teletext operations, and Chapter Eight discusses miscellaneous services that can be found on satellites. Chapter Eight also has a complete detailed section on the Hughes C-band domestic satellite facilities.

The information contained in this book is not intended to encourage the misuse of satellite services. Rather it has been prepared by the authors as a primer on what is available there. *FCC rules and regulations do not prohibit SWL'ing. They do prohibit selling or deriving a commercial benefit from what is received.* The ultimate responsibility lies with the user.

In some areas, technical descriptions in this book are not exactly what the engineer may want. However, the average TVRO owner, TV technician, or beginning TVRO dealer will find this book to be a good reference manual to have on the shelf.

The Hidden Signals on Satellite TV is published by Universal Electronics, 4555 Grove Road, Suite 34, Columbus Ohio 43232. Copies are available from Ham Radio's Bookstore, Greenville, NH 03048, $14.95 plus $3.50 shipping and handling.

N1ACH



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flea market

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WANTED: Old Crosley Radio Model 50, 51, 52, & "Pup". K4NBN "No Bad News"

TUBES: Selling 40 year collection. Advise requirements for quotes. Desperately need manuals for Hallicrafters SX122A. M. Levy, 4141 Krupp Drive, El Paso, Texas 79902. Apt. One.

ATTENTION Builders: For sale. Two 4CX250R/7580W tubes with silver bypass sockets (unused), low profile plate tran. 2000 v.c.t., 1/2 a., filament tran., high cap. filters, bridge rectifier. $275 prepaid UPS. Postal money order only. R.E. Peterson, W0PQX, 1302 S. Arlington Ave., Duluth, MN 55811

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TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details. W8UUR, 1025 Wildwood Road, Quincy, MI 49082.

NEW VLF CONVERTER by K1RGO covers 2 kHz to 500 kHz. AM broadcast rejection > 100 dB at 1 MHz, i-f rejection 130 dB. 3.5-4.0 MHz (L-101/80) or 4.0-4.5 MHz (L-101/70) i-f tuning available. $49.00 postpaid cont. US. Free bochure. LF Engineering Co., 17 Jeffry Rd., East Haven, CT 06512.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

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ATLAS 350XL owners group. Send QSL card with s/n your rig. Know anyone who repairs them? Have any technical information to share? Any questions? Rod, N5NM, Box 2169, Santa Fe, NM 87504.

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SCHEMATICS: Radio receivers 1920/60's. Send name brand, model, SASE. Scaramella, P.O. Box 1, Woonsocket, RI 02895-0001 (602) 897-2534

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TECHNICAL BOOKS. "Electronics with Digital and Analog Integrated Circuits" by Richard Higgens of the University of Oregon/Eugene. Combines the best of the "electronics for scientists" and the "hobbyist" or "how to" approach. Circuit principles and "golden rules", which won't become obsolete with envolving technology, are emphasized. $30.95 + 3.50 shipping. "Digital Design" by Morris Mano of California State University. Presents the basic tools used in the design of digital circuits. $33.95 + 3.50 shipping. Check or money order. Technical Advice, Inc., Box 4757, Englewood, CO 80155.

CUSTOM embroidered emblems — Cloisonne enameled pins, your design, low minimum, excellent quality, free booklet. A.T. Patch Co., Dept. 65, Littleton, NH 03561. (603) 444-3423

HAM RADIO Magazine collection. Bound volumes 1972, 1973, 1974 and 1975. In HR binders 1976, 1977, 1978 and 1979. $120 for lot plus UPS shipping. US only. W4UCH, Box 1065, Chautauqua, New York 14722 (716) 753-2654.

WANTED: Cash paid for used speed radar equipment. Write or call: Brian R. Esterman, PO Box 8141, Northfield, Illinois 60093 (312) 251-8901

WANTED: Old microphones, remote mixers other misc related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

WANTED: Used microwave equipment. D. Mitchell, K8URL, 1 Cider Mill Lane, Upton, MA 01568. (617) 529-4638.

RECONDITIONED TEST EQUIPMENT $1.00 for catalog. Walter, 2697 Nickel, San Pablo, CA 94806.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

"HAMS FOR CHRIST." Reach other Hams with a gospel tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786

WANTED: Early Hallicrafter "Skyriders" and "Super Skyriders" with silver panels, also "Skynder Commercial", early transmitters such as HT-1, HT-2, HT-8, and other Hallicrafter gear, parts, accessories, manuals. Chuck Dachis, WD5EOG, The Hallicrafter Collector, 4500 Russell Drive, Austin, Texas 78745.

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ELECTRON TUBES: Receiving, transmitting, microwave... all types available. Large stock. Next day delivery most cases. Daily Electronics, PO Box 5029, Compton, CA 90224 (213) 774-1255

COMING EVENTS

Activities — "Places to go . . ."

MICHIGAN: The Cherryland Amateur Radio Club's 12th annual Swap 'N Shop, Saturday, February 9, Immaculate Conception Middle School Gymnasium, 218 Vine Street, Traverse City. 9 AM to 2:30 PM. General admission $2.50. Single tables $3.00. Talk in on 146.85 and 146.52 simplex. For information: Paul Nepote, KA8HIB, Chairman, 802 Fern St., Traverse City, MI 49684. Please SASE.

FLORIDA: The Fort Myers City of Palms ARC's annual Hamfest, Saturday, February 23, Moose Hall, 1900 Park Meadow Drive. ARRL volunteer license exams will be given by previous registration only, no walk-ins. Indoor flea market tables $10.00. Admission $3.00 at door, no mail orders. Ample free parking. Talk in on 146.28-146.88

OHIO: Dayton Hamvention, April 26, 27, 28, Hara Arena and Exhibition Center, Dayton. Admission $8 advance, $10 at door. Good for all three days. Banquet $14 advance, $16 at door. Flea market space $17 in advance for all three days. Technical, ARRL and FCC forums. New products and exhibits. Special group meetings. YL forum. International VHF/UHF conference. Amateur of the Year Award. Special achievement awards. Pre-registration starts January 1, 1985. For further information; Dayton Amateur Radio Association, Box 44, Dayton, OH 45401 or phone (513) 433-7720.

MASSACHUSETTS: The Mount Tom Amateur Repeater Association will host its first annual indoor Flea Market, March 3, 9 AM to 4 PM, Knights of Columbus Hall, Elder Council 69, Granby Road, Chicopee. General admission $1.00. Kids and spouse free. Tables $8.00 door, $7.00 advance. Set up 8-9 AM. Food and drink. Contact Mickey Yale, N1CDR, 6 Laurel Terrace, Westfield, MA 01085. (413) 562-1027.

OHIO: Teays Amateur Radio Club's 7th annual Hamfest, March 3, K.C. Lodge, Circleville. Sellers set up 6 AM. Open to public 8 AM. Advance tickets $2.00, at door $3.00. 8' tables $4.00 advance, $5.00 at door. For table reservations SASE to Joe Subich, AD8I, 7825 State Rt. #188, Circleville, Ohio 43113. Talk in on Circleville repeater 147.18. For information: Len Campbell, WB8PPH, 8951 State Rt. #188, Circleville, Ohio 43113.

MISSOURI: The Jefferson Barracks Amateur Radio Club's 25th annual Amateur Radio Auction, Friday night, March 8, St. Louis Firefighters Hall, 5856 Gravois at Christy, in south St. Louis City. Starting time 7:30 PM. This is the Silver Anniversary of this event and the club is going all out to make this a memorable event. For information: Jefferson Barracks ARC, c/o Carl H. Hohenberger, W0BZP, 5266 Parker Avenue, St. Louis, MO 63139.

WISCONSIN: The Ozaukee Radio Club's 7th annual Swapfest, Saturday, May 4, 8 AM to 1 PM, Circle B Recreation Center, Highway 60, Cedarburg. Admission $2.00 in advance, $3.00 at the door. 4' tables $2.00 each advance only. Food and refreshments available. Sellers admitted at 7 AM for setup. For tickets, tables, maps or information send business SASE to 1985 ORC Swapfest, 101 E. Clay St., Saukville, WI 53080.

VIRGINIA: The Vienna Wireless Society's annual Winterfest, Sunday, February 24, Vienna Community Center, 120 Cherry Street, Vienna. Admission $4.00. Doors open at 8 AM. For vendor and tailgate applications SASE to Earl Hohbein, N4FSW, 4602 Lawn Court, Fairfax, VA 22032. Coffee and food available. Talk in via NVFMA 146.31/91 and VWS 146.085.685 and 147.51 simplex. For information: Vienna Wireless Society, PO Box 418, Vienna, VA 22180.

WISCONSIN: The Tri-County Amateur Radio Club's annual Hamfest, March 17, 8 AM to 3 PM, Jefferson County Fairgrounds, Jefferson. Tickets $2.50 advance, $3.00 at door. Tables $3.00 advance, $4.00 at door. Plenty of food and free parking. Doors open at 7 AM for sellers only. Talk in on 144.89/145.49, 146.22/146.82, 146.52/146.52. For information, tickets and tables SASE to Bob Barker, K9RIJ, 724 Burdick, Milton, WI 53563.

MARYLAND: BARC, the Baltimore Amateur Radio Club's 1985 Greater Baltimore Hamboree and computerfest, March 31, Maryland State Fairgrounds Exhibition Complex, Timonium. Gates open 8 AM. Admission $4.00, children under 12 free. Amateur radio, personal computer and small business computer dealers, guest speakers, large, hard surface outdoor tailgate area. Food service, free parking. Overnight accommodations available in immediate area. For information and table reservations GBH & C, PO Box 95, Timonium, MD 21093-0095. (310) 561-1282.

TEXAS: The Midland Amateur Radio Club's annual St. Patrick's Swapfest, Saturday, March 16, 10 AM to 5 PM and Sunday, March 17, 8 AM to 2:30 PM, Midland County Exhibit Building, north of Highway 80. Pre-registration $5, $6 at the door. Tables $6 each. Refreshments and food available. Volunteer examiner tests for all categories. For information and reservations: Midland ARC, PO Box 4401, Midland, TX 79704.

WASHINGTON: The Mike and Key Amateur Radio Club's 4th annual Electronic Flea Market, March 9, 9 AM to 7 PM, Western Washington Fairgrounds, Puyallup. Admission $2.00 Spouse and kids free. Flea market tables $15.00. Consignments 10%. Dealers, exhibitors and demonstrations. Satellite station. Packet station. Super snack bar. Free parking. For reservations: Electronic Flea Market, 20903 NE 77th, Redmond, WA 98052. (206) 883-3012.

MICHIGAN: The 2nd annual Amateur Radio Auction sponsored by the Holland Amateur Radio Club, Saturday, March 9, Hudsonville High School Auditorium, 5051 32nd Avenue, Hudsonville. No admission fee. Equipment can be checked in from 8 AM to 12 noon. Auction 9 AM to 1 PM. A 10% donation for each item sold. Refreshments available. Talk in on 146.06 and 62. For information: Dan Ruiter, KC8KN, 7106 Michael Drive, Hudsonville, MI 49426.

MINNESOTA: The Robbinsdale Amateur Radio Club's 4th annual Midwinter Madness Hobby Electronics Snow, February 23, 8 AM to 2 PM, Totino-Grace High School, 1350 Gardena Avenue NE, Fridley (suburb of Minneapolis). Admission $4 at door. Manufacturers, dealers, flea market. Talk in on 147.60/00 K0LTC Repeater and 146.52 simplex. For information: Robbinsdale ARC, PO Box 22613, Robbinsdale, MN 55422 or call Bob (612) 533-7354. All Amateur Radio tests will be given. For information: Elmo Nygard, 4151 Adair Avenue N, Robbinsdale, MN 55422.

INDIANA: The LaPorte ARC's winter Hamfest, Sunday, February 24, at the LaPorte Civic Auditorium. Donation $2.50. Tables $2.00 advance, $2.50 at door. Talk in on 52 simplex. For information and reservations: LARC, PO Box 30, LaPorte, IN 46350.

MICHIGAN: The 15th annual Livonia Amateur Radio Club's Swap 'n Shop, Sunday, February 24, 8 AM to 4 PM, Churchill High School, Livonia. Plenty of tables, refreshments and free parking. Talk in on 144.75/5.35 and 52 simplex. For information or table reservations SASE to Neil Coffin, WA8GWL, c/o Livonia ARC, PO Box 2111, Livonia, MI 48151.

OREGON: The 5th annual Salem Mini-Hamfair, February 23, Polk County Fairgrounds. Seminars, commercial displays, Amateur license exams and a large flea market. Admission $4.00. Flea Market set up 8 AM. Doors open 9 AM. Talk in on 146.25/.86 and 146.52. For information: Sales Repeater Association, PO Box 784, Salem, OR 97308.

KENTUCKY: Annual Glasgow Swapfest, Saturday, February 23, 8 AM till ????? Glasgow Flea Market Building, 2 miles south of Glasgow off 31E. Free coffee, free parking, large flea market and a friendly gathering of hams. Admission $2.00. No extra charge for exhibitors. One free table per exhibitor. Extra tables $3.00 each. Talk in on 146.34/94 or 147.63/03. For information: N4HCO, Rt. #4, Box 354, Glasgow, KY 41241.

OHIO: The Cuyahoga Falls Amateur Radio Club's 31st annual Electronic Equipment Auction and Hamfest, Sunday, February 24, North High School, Akron. 8 AM to 3 PM. Tickets $3.00 at the door. Sellers may bring own tables or some will be available to rent. Talk in on 87/27. For information or reservations SASE to Bill Sovinsky, K8JSL, 2305 — 24th St., Cuyahoga Falls, Ohio 44223 or call (216) 923-3830.

OHIO: The 24th annual Mansfield Mid-Winter Hamfest/Auction, Sunday, February 10, Richland County Fairgrounds, Mansfield. Doors open to public 8 AM. Tickets $3.00 advance, $4.00 at door. Tables $5.00 advance, $6.00 door. Hall tables available. Talk in on 146.34/94. For information, advanced tickets/tables, SASE to Dean Wrasse, KB8MG, 1094 Beal Road, Mansfield, Ohio 44905. (419) 589-2415. An ARRL/VEC license exam will be held at Mansfield campus, Ohio State University, North Central Technical College at 1 PM day of the Hamfest. SASE with 610 and check for $4 payable to ARRL/VEC to: Lloyd Nelson, N8BAZ, 630 Oak St., Lot 82, Mansfield, Ohio 44907.

Operating Events — "Things to do . . ."

FEBRUARY 1-10: The North Okanagan Radio Amateur Club will operate a special station to commemorate the 25th anniversary of the Vernon Winter Carnival, the largest in Western Canada. Listen for VE7NOR, our club station, from 2100Z to 2400Z daily. Frequencies 28.525, 21.375 and 14.225. For a special certificate send $1.00 or 2 IRC's to cover postage along with QSL information to: Box 1706, Vernon, BC, Canada V1T 7T9.

MARCH 9 and 10: West Coast 160 Bulletin CW Contest. 0000 GMT 3/9 to 2359 GMT 3/10. Single operators only. Exchange RST, QTH. Subscribers, non subscribers. Scoring: 10 points per QSO. Multipliers states, VE prov., country. Log info date, time rs(t), QTH. Send logs to R. Koziomkowksi, 5 Watson Drive, Portsmouth, RI 02871. Must be postmarked before April 30, 1985.

MARCH 16 and 17: DARC International SSTV Contest. 1200 UTC 3/16 to 1200 UTC 3/17. All for SSTV authorized bands. Exchange only two way video count. Exchange call sign signal report and QSO serial number on video. Logs must indicate date, time (UTC), call sign of station worked and complete exchange for each valid QSO. Name and address of entrants. Mail contest logs to Heinz Moestl, DE8BUS, PO Box 1123, D6473 Gedern 1, W. Germany.

THE GUERRI REPORT

Ernie Guerri W6MGI

superchips come of age

About five years ago the Department of Defense embarked on an ambitious program to develop a family of Very High Speed Integrated Circuits (VHSIC). The objective was to design a group of VLSI chips with 1.25 micron features and a computational throughout of 5 × 10^{11} gate-Hertz. The program has cost over $350 million, but has yielded some impressive results. Signal processing chips capable of performing 100 million multiplications per second, memories with densities of 5 million transistors per square inch, a complete vector arithmetic processor and a self-contained convolutional decoder (a form of detector that can recognize very complex codes without going through all the arithmetic) have now been implemented.

Although some of the chips being fabricated have specialized military functions, most are designed to form a family of arithmetic and signal processing functions. Both bipolar and CMOS designs have been implemented, and some include current mode logic, which can perform subnanosecond operations. All of this exciting work is intended to be available eventually for commercial use. The next phase of the DoD program will aim at 0.5 micron features and nearly 100 times the speed — by the end of this decade!

advanced materials serve RF needs

We sometimes forget the key role that proper materials selection plays in the correct performance of many products. For example, we all know that a stable oscillator requires a tuned circuit with a high Q; that the active element must have carefully controlled gain and phase characteristics; and that all these parameters change with temperature. In the region below 100 MHz, the most stable oscillators have been traditionally built using silicon dioxide (quartz) crystal elements as the resonator. In the VHF and microwave region, cavity resonators have been the principal alternative.

The ideal resonator would have a very high Q, zero temperature coefficient, high permittivity (low radiation losses) and be supported by a dielectric with no electrical or thermal losses. Designers of semiconductor oscillators find series resonant (low impedance) tank circuits convenient. This means that a high Q series tank will have high capacity and low equivalent inductance. The modern alchemists have been at work on the problem and have now given us nearly perfect materials with which to solve the problem. Dielectric resonators made of ceramic materials with high dielectric constants (over 30), have made possible microwave resonators with unloaded Q's of 25,000 at 4 GHz and over 10,000 at 12 GHz. These materials have extremely low losses and excellent temperature characteristics, making possible the production of oscillators offering crystal-like performance at microwave frequencies

But hold on to your hat! In the search for the best dielectric materials to support these resonators, the thermal tiles used to protect the Space Shuttle on re-entry emerged as an optimum material. With a dielectric constant of 1.15 and legendary thermal characteristics, the tiles, made of foamed quartz, become a nearly perfect enclosure/support for a dielectric resonator. Using this combination of advanced materials, a major defense contractor has fabricated an oscillator that exhibits only ±10 kHz drift per month — at over 6 GHz!

Gallium arsenide digital circuits are already available in the 4 GHz region, so it shouldn't be too long before we can have phase locked loops at 5 GHz or so. Combined with SAW or dielectric resonator filters, image-free upconversion receivers through 2300 MHz can be a reality for Amateurs of the next decade.

telephones to be more versatile

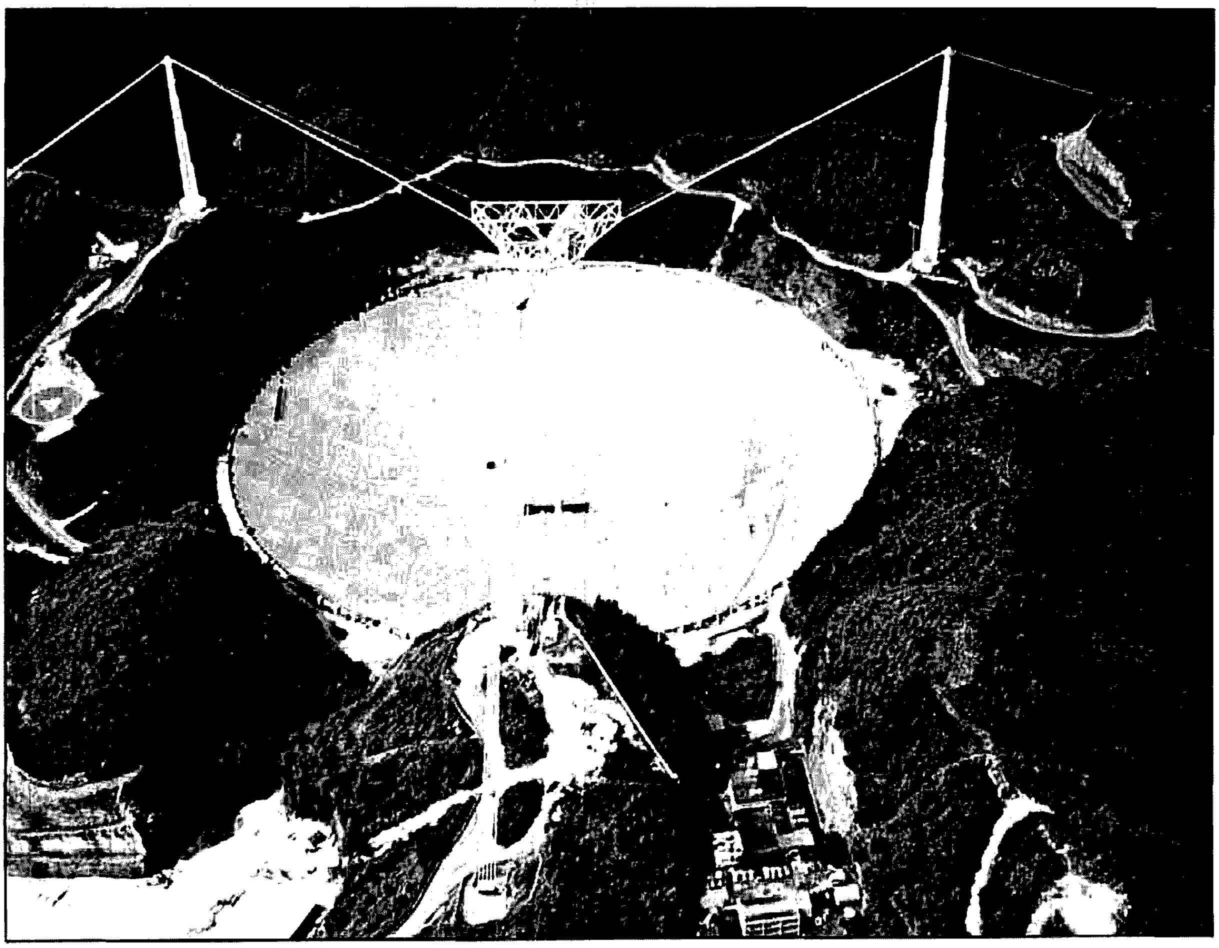
Over the past few years the plain old family telephone has become a hot consumer product. We've seen a proliferation of telephones that offer just about every convenience we could want in a single desk instrument. Up to 60 memories, on-hook dialing, full duplex speakerphone, as well as time and message unit recording, are now available in a multitude of styles at competitive prices. All of this capability has been made possible by the development of specialized ICs which have considerable processing power in their structure.

Most telephones now have 12-digit keypads, with each digit identified by a unique tone. If each digit is thought of as a single data "bit," then our ordinary keypad becomes a 12-bit code

generator. Twelve bits equals 2^{12}, or 4096 possible codes if we use each tone just once in a 12-digit binary number. If your telephone were used as a small programmable data instrument, then you would have more than enough codes, and ample data resolution, to control most of the common household items that now each have individual and far less accurate, controllers. Household heating and cooling, selection of cooking times and temperatures, cost-effective regulation of hot water use and temperature, security entry codes, and choice of entertainment channels, for example, can all be controlled, changed, and recorded on our ordinary telephone. Moreover, because the telephone is connected to the rest of the world, we can change this household data from remote locations as the need arises. For Amateurs, the possibilities are especially appealing. Many repeaters now have their functions controlled by tone decoders, in addition to having all the usual auto-patch facilities. Antenna rotor controllers that accept bearing instructions from the telephone *touchpad, remote frequency setting of* transceivers, and a variety of data management functions will all become much easier and require only a single instrument for control. Someday we may even be able to reach out and . . . well, you know.

ham radio

ham radio magazine



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focus on communications technology

AMATEUR RADIO AND THE SEARCH FOR EXTRATERRESTRIAL INTELLIGENCE • *designing low-voltage power supplies* • *harmonic signal mixer* • *controlled vertical radiation rhombics: part one* • *get on 6 meters — inexpensively* • *fast-scan ATV amplifier* • *predicting solar outages* • *W1JR, W6SAI, KØRYW, and THE GUERRI REPORT*

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magazine

MARCH 1985

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contents

REFLECTIONS

what's wrong with Amateur Radio?

A lot of fuss is being made these days about the future of Amateur Radio. The number of new hams entering the hobby is down significantly; 10.2 percent fewer licenses were granted in 1984 than in 1983. The number of hams who upgraded in 1984 was down 30 percent from 1983, and of the hams currrently licensed, 25 percent will choose not to renew their tickets. For the first time in ten years there was, at year's end, a net loss of licensed Radio Amateurs.

What does this all mean?

Very simply, it means that Amateur Radio has some big problems.

Not all the problems are within. There's a lot of pressure from outside the hobby. Municipalities seek to limit or prohibit tower construction and control sources of possible RFI. Commercial and other interests want our spectrum space. The FCC, on one hand, has made getting the Novice license easier; on the other hand, upgrading has become more difficult — though perhaps this is no more than a temporary effect of implementing the Volunteer Examiner program.

It all adds up to an unhealthy situation. And unless all of us — each and every ham in the United States — gets involved, very soon, Amateur Radio as we know it *may cease to exist* within the next ten years.

Some might say this is an overly provocative statement, and that I'm seriously exaggerating the situation. I don't think so.

A few days ago I read an editorial by Al Dorhoffer, K2EEK, in the February, 1985, issue of *CQ*, and spoke with Al by phone later.

While *ham radio* and *CQ* address two uniquely different segments of Amateur Radio, our futures are irrevocably linked to the future of this hobby. In his editorial, Al pointed out that we need to involve our children in Amateur Radio. There's a dual meaning here, because by "our children," Al means both our own offspring and children in general. Before you go out and rail against what is or isn't being done by everybody else, let me ask you to look at your own household. On close examination, I think you'll find some interesting information there.

If you do have kids at home, are they hams? (If they are, congratulations. You can skip the next paragraph.)

If they aren't hams, why aren't they? Have you done everything possible to interest them in Amateur Radio? Or have you bored them with statistics and failed to show them how much fun Amateur Radio really can be?

Watching Mom or Dad on the air isn't the answer.

Hands-on experience is.

Studies in group dynamics show that in any group, 90 percent of the work is done by 10 percent of the people involved. If you're sitting back waiting for somebody else to interest *your* kids — or your neighbor's — in Amateur Radio, you can sit back and watch Amateur Radio wither away. All across the country there are local pockets of enthusiasm that seem to turn out the majority of new hams year after year. If your area isn't one of these, what are you doing to make it so?

Examining the latest figures on VEC exams given nationwide, I was discouraged to find that while some areas are being well served, with aggressive recruiting programs and regularly scheduled examinations, others were still waiting for their first exam sessions.

Another point K2EEK mentioned must be emphasized: Amateur Radio is not a private club. Once we've attracted new hams — whether they're young people or adults — we owe them all the nurturing, all the help, we can give them. When questions are asked, help should be given. Several months ago we received a letter from an irate newcomer to the hobby. An eager Novice, he'd joined a local ham club only to be totally ignored by most of the membership. Meetings were spent discussing ways to beat the cable company and how so-and-so was such a ding-a-ling. Nobody was interested in helping him enter the mainstream of Amateur Radio.

On the way to visit relatives over Christmas, I was talking with a friend on a repeater in central Pennsylvania. After we signed, two other hams got on and proceeded to have a donnybrook. If I'd been trying to demonstrate the wonders of Amateur Radio to a friend at the time, I would have been terribly embarrassed. What would *he* have thought?

The bottom line is that before we go blaming anyone else for the problems in Amateur Radio, we have to look at ourselves first. If you have kids at home, and they're not hams, *how come*? If your club doesn't have any young members, *why not*? And if you're quarrelling on the air, *what are you telling prospective hams about Amateur Radio*?

It's up to *us* to make sure that Amateur Radio survives. If we do nothing but blame everyone else, it won't — it's really that simple.

Physician, heal thyself.

J. Craig Clark, Jr., N1ACH
Assistant Publisher

presstop de W9JUV

SENATE SUPPORT FOR PRB-1, ARRL'S PETITION REQUESTING FCC PREEMPTION of state and local restriction of Amateur antennas and activities, is being sought by Senator Barry Goldwater, K7UGA. In his Senate Resolution 36 he states it is the "sense of the Senate" that the FCC affirm that state and local governments must not pass laws that "discriminate unreasonably among Amateur Radio Antennas..." or "have the effect of prohibiting or frustrating the transmission or reception of Amateur Radio communications...." A parallel proposal, Senate Resolution 35, extends the same protection to TVRO satellite antennas.

If The Senate Does Vote In Favor Of Resolution 36 it will provide a big boost for PRB-1, even though it is not binding on the FCC, Amateurs everywhere should contact their senators immediately to urge them to support Senate Resolution 36—and also Senate Bill 66, Goldwater's new bill making intentional interference to any radio service a federal crime.

AMATEUR VHF/UHF FREQUENCY COORDINATION SHOULD BE A NATIONWIDE, organized effort, the FCC has proposed in a far-reaching NPRM that's just been released. An increasing number of requests from Amateurs for FCC assistance in settling repeater conflicts, plus the general awareness that the problem will continue to worsen with ever-increasing VHF/UHF band usage, is behind the FCC's proposal. One example is the recent case of a New England repeater operator who sued another repeater group and the council that coordinated them on his machine's frequency. That conflict was resolved only with the intervention of top ARRL officials, and the second machine is now looking for another frequency.

Role Model For An Amateur Radio National Volunteer Coordinator is the land mobile industry, where industry trade groups have set up a coordination effort that is highly respected and considered very effective. For Amateur Radio, the ARRL is the logical—if not the only—organization capable of taking on an effort of such magnitude. However, in the past the League has steadfastly maintained its distance from coordination efforts, limiting its involvement to band plan generation and publishing its repeater directory. In its 1985-86 Repeater Directory, however, the League does plan to indicate which repeaters are "coordinated" and which are not. The actual coordination has been done by state or regional councils or individual volunteer coordinators, with varying degrees of success.

The Comment Period On The National Coordinator Proposal, designated PR Docket 85-22, has been made unusually long. Comments are not due until July 1, while Reply Comments can be submitted until September 30.

The Midwest Meeting To Discuss 20 kHz Spacing On 2 Meters' High End was considered very successful, despite record-breaking cold over the January 19-20 weekend. About 40 people—representing Ohio, Indiana, Kentucky, Wisconsin, Ontario, and of course Michigan, whose decision to shift to 20 kHz next year had precipitated the meeting—were present. Tone of the ARRL-sponsored session was generally positive, but with little apparent enthusiasm outside of Michigan for the change. The meeting lasted over five hours, and when it ended the Michigan representatives agreed to take the matter back to their council for further discussion. A followup meeting may take place at the Dayton Hamvention in April.

Texas' Decision Whether Or Not To Adopt 20 kHz Spacing was expected at the February 16 meeting of the Texas VHF FM Society. Nebraska Coordinator WAØWRI reports his state is not considering a shift to 20 kHz spacing, contradicting February Presstop. He also says there's little support for a move in Iowa, and he expects it to be voted down in Kansas.

STIMULATING INTEREST IN AND GROWTH FOR AMATEUR RADIO was to be the subject of a closed, all-day meeting January 31 just before the Miami Tropical Hamboree. Initial idea for the meeting came from Mike Lamb, N7ML, of AEA, and it was picked up by many other Amateur Radio manufacturers, distributors, and publishers. Focus of the discussion was to be on how Amateur Radio, whose growth has stagnated in recent years, can be revitalized, and what can be done to make the ARRL's growth targets for the next few years possible. A report on this crucial meeting will appear in ham radio's April issue.

A VEC NO LONGER HAS TO ADVISE THE FCC FIELD OFFICE OF FORTHCOMING EXAMS under an order adopted by the Commission in late January. It should be effective before this reaches print. Another rules relaxation to simplify VE paperwork, also adopted, is due shortly.

The 30-Day Wait Before Taking A Failed Exam May Be Dropped altogether under a Notice of Proposed Rule Making released in January. The Commission's proposal, PR Docket 85-21, would remove the mandatory delay from the exam rules; an applicant, if he wished and the VEs permitted, could retake a failed exam the same day he flunked it! The only limitation would be that the VEs would have to give him a different set of questions for the second attempt. In addition, under the proposed rules change a VEC could, if it wished, set a waiting period of its own choosing.

Comments On PR Docket 85-21 Will Be Due About April 1, with Reply Comments two weeks later. (The exact dates had not yet been set at press time.)

IMPLEMENTATION OF WARC CHANGES IN MICROWAVE AMATEUR BANDS have been proposed in yet another FCC NPRM. The most significant change in the proposal is probably the loss of 25 MHz at the bottom of the 21-cm band, making it 1240-1300 MHz instead of 1215-1300 MHz.

Comments On The WARC Microwave Implementation Proposal, PR Docket 85-23, will be due about the first of April. Reply Comments will be due in mid-April.

new trends in communication technologies: radio astronomy and the search for extraterrestrial intelligence

"Man's first step toward maturity may be to contact life beyond the solar system."

—Bernard M. Oliver

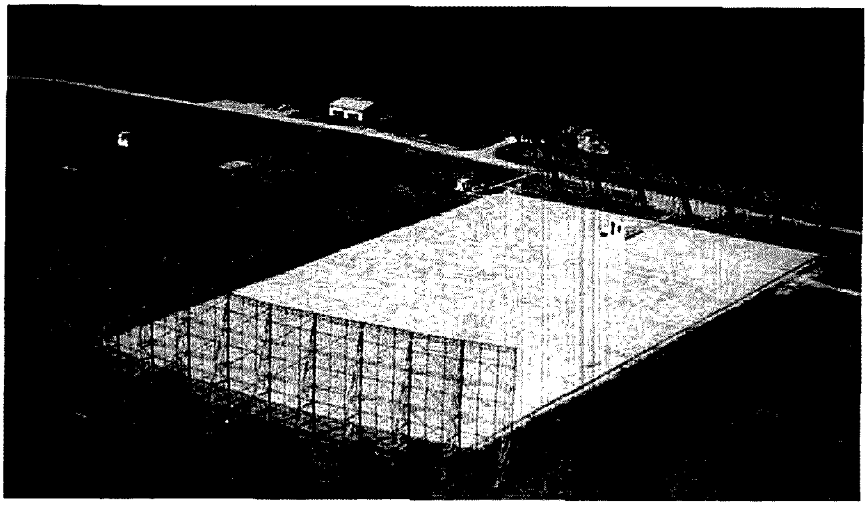


fig. 1. This radio telescope — larger than three football fields — allows for the detection of thousands of radio sources extending to distances of 10 billion light-years (1 light-year = 6 × 10^{12} miles). *(Photo courtesy of Ohio State University Radio Observatory.)*

If you're looking for a new technical challenge, you may find SETI, the Search for Extraterrestrial Intelligence, to be just the frontier you've been seeking. With today's microwave technology, it is possible to communicate anywhere within our galaxy. And although radio astronomy is still a relatively young science*, Amateur Radio operators have access to most of the state-of-the-art components found in a professional radio astronomy center, with the possible exception of the very large antennas. Because nobody knows what the first extraterrestrial signal will be like, there is ample room for ham ingenuity. After all, if hams were to be the first to communicate with extraterrestrials, it wouldn't be the first time a major scientific breakthrough had been made by hams — remember, not too long ago hams discovered the ionosphere.[1]

For several years I have been contemplating the construction of a system that would allow the reception of intelligent information generated by a hypothetical 1-Gigawatt EIRP (real power times antenna gain) transmitter located approximately 25 light-years away. I prepared this article in order to share some of the knowledge gained during this process and to provide a comprehensive overview of recent progress in radio astronomy (including SETI) and to assess what Amateurs can do with even limited resources.

*Unlike other events in science, the birth of radio astronomy can be traced precisely — to the early 1930's when Carl B. Jansky, a Bell Telephone radio engineer, performed antenna noise studies for long-range communications at the wavelength of 14.6 meters. With these studies, Jansky proved that extraterrestrial radiation can be received. Jansky's experiments were followed, in the late 1930's, by Grote Reber, W9GFZ, an amateur astronomer who designed and built the first parabolic radiotelescope and performed a survey of the galaxy at the wavelength of 1.9 meters.

By Cornell Drentea, WB3JZO, 7140 Colorado Avenue North, Brooklyn Park, Minnesota 55429

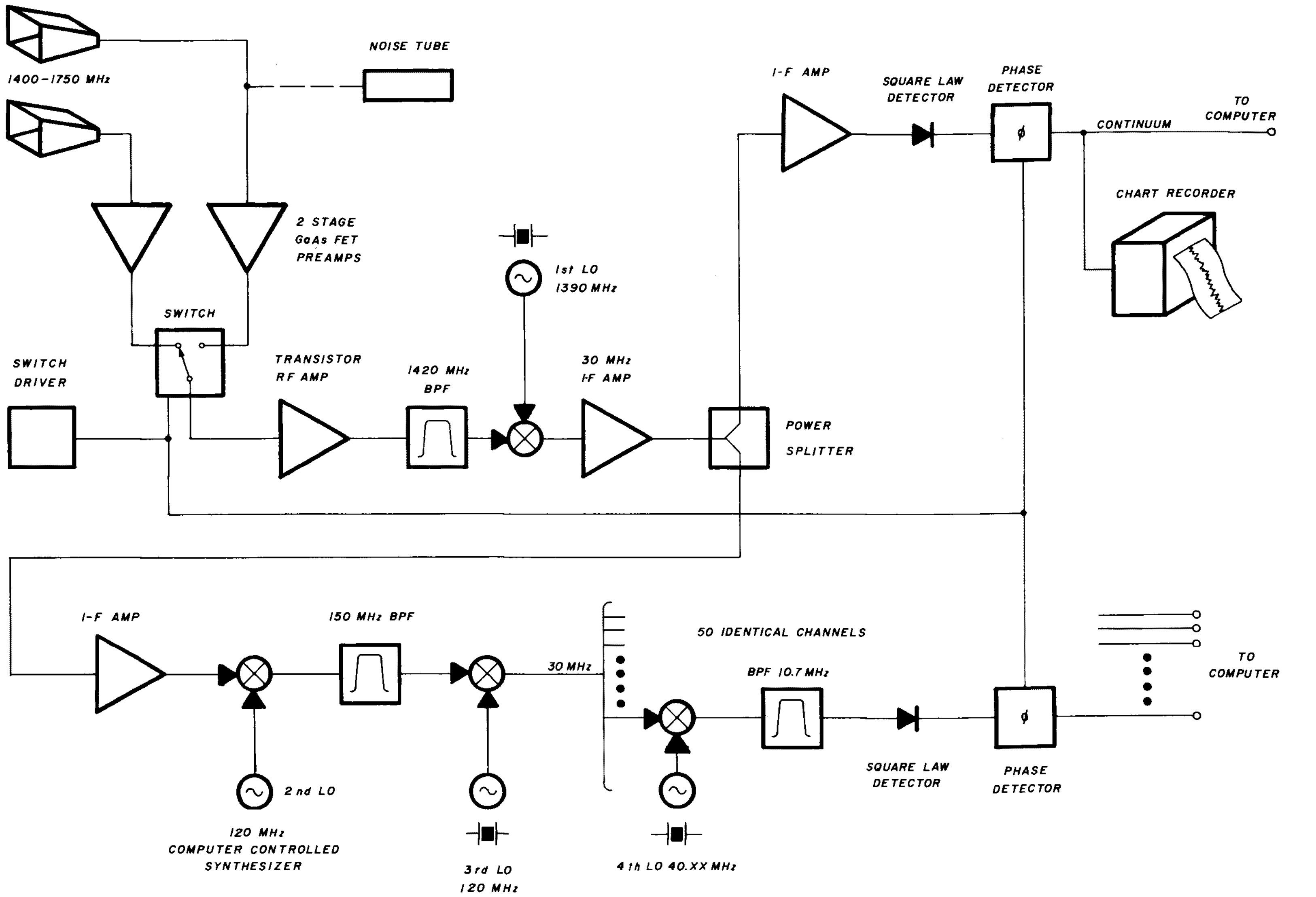


fig. 1B. Block diagram of the Ohio State University radio telescope configured for SETI. The system covers an instantaneous bandwidth of 500 kHz through a bank of 50-channel IF filters at 10.7 MHz (individual filter bandwidth is 10 kHz). (*Courtesy of Ohio State University Radio Observatory.*)

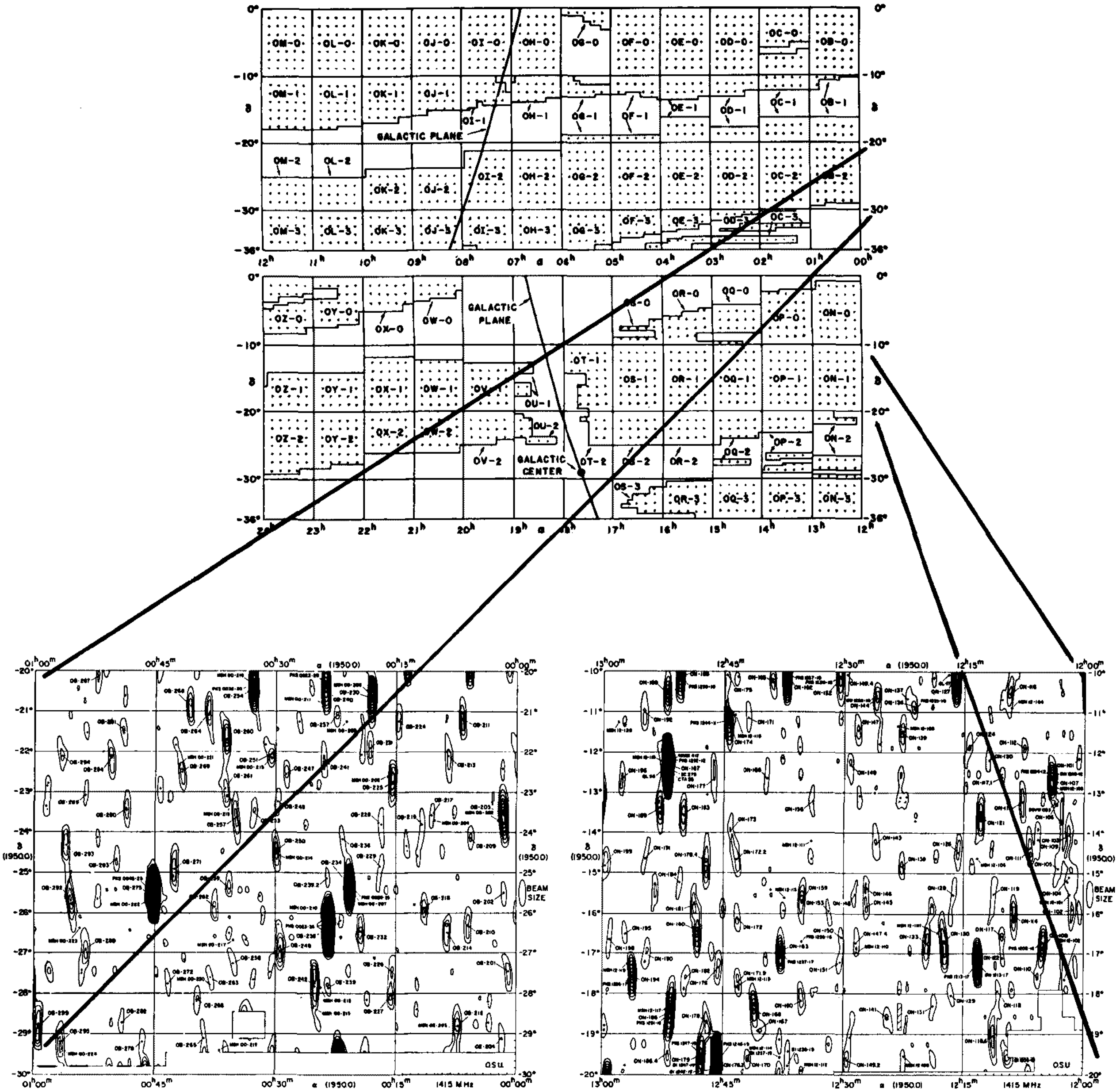


fig. 2. Master map of an Ohio University radiometer search. Maps below illustrate details of two sections indicated above. (*Courtesy of Ohio State University Radio Observatory.*)

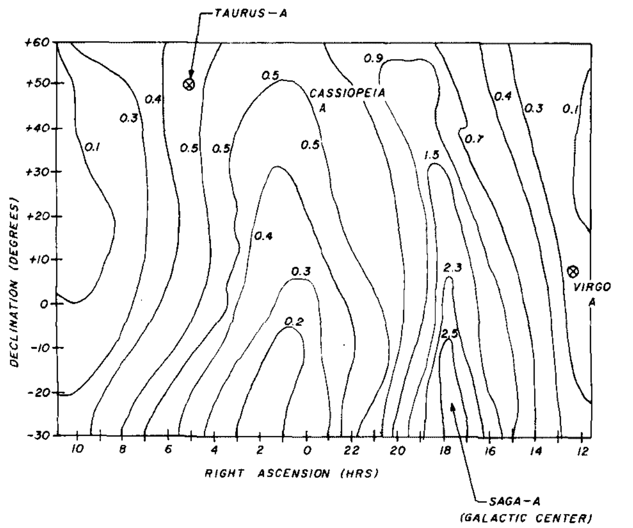


fig. 3. A map of the radio sky obtained with a simple amateur radiometer. Because of the antenna's low resolution, discrete radio sources appear as a set of signal gradient lines resembling a surveyor's map.

conventional radio astronomy

There are two current trends in radio astronomy. The first, and by far most popular, involves the study of wide-band noise generated by powerful sources within our galaxy or in other galaxies. The second, which occupies only a small fraction of the total activity, employs extremely narrow bandwidth receivers designed for the detection of intelligent monochromatic signals in the microwave regions of the frequency spectrum where the level of intergalactic noise is lowest.

Within the context of the first trend, it is relatively easy for an Amateur to build a radiometer receiver intended for casual observations of very strong radio sources. Because of the uniform distribution of wide-band noise over the receiver's bandwidth, no particular attention to local oscillator stability is required. There would likewise be no need for precise tuning to compensate for the Doppler shift in the incoming signals caused by Earth's rotation and by the relative motion between the observed celestial object and our own solar system.

Professional radiometers employ giant steerable antenna arrays that allow for the detection of natural radio sources located at great distances from Earth. A continuum survey of the sky was made by the Ohio University Radio Observatory, using the installation shown in **fig. 1A** and **1B**. The receiver employed a liquid nitrogen-cooled parametric amplifier with a calculated system temperature of 95 degrees K. The bandwidth was 8 MHz and the output was integrated over a 10-second period. Concurrent recording was performed after processing the data through IBM 7094 and 1620 computers. The entire system was synchronized with a sidereal clock accurate to within 0.05 second. Results have been plotted in maps of the region surveyed as shown in **fig. 2**. In its search of almost the entire sky, from −36 to +63 degrees, the Ohio State project found 20,000 radio sources.

While such performance cannot be duplicated by the backyard radio astronomer, remarkable results can be obtained with relatively modest installations. An Amateur radiometer is usually a high-gain VHF/UHF superheterodyne receiver that features simple amplitude modulation detection followed by a DC amplifier equipped with an integrator. The output transducer can take the form of a conventional chart recorder or some other measuring device, or can be an analog-to-digital (A/D) converter connected to a microcomputer using a dot matrix printer for the output. The format would be digitized flux samples (values from 0V to 9V) at, for example, 1-second intervals printed out in 60-second columns for a total of 1 hour of information per page. With such a receiver and a multi-element beam antenna — and with considerable skill and patience — a serious Amateur can map the radio sky in a short time.

The methodology employed involves pointing the antenna at a known celestial location and then relying on the Earth's rotation to bring in the various natural radio sources. This requires knowing celestial coordinates and times as well as converting the recorded information and antenna position into the right ascension and declination values in order to plot the signal onto a celestial map that would resemble the actual sky (see **fig. 3**).

A typical multi-element beam antenna with a major lobe beamwidth between the half power points of approximately 30 degrees would allow a natural radio source to pass through its beam in approximately two hours (the apparent rate of movement of a celestial object is 15 degrees per hour at the equator). This in turn would be sufficient to allow for the reception of strong Milky Way sources such as Cygnus A, located approximately 500 light-years away, and Cassiopeia A, located approximately 200 light-years away, regardless of the system's bandwidth or operating frequency.

very-long-baseline interferometry (VLBI)

In order to increase the resolution of a simple radiometer so that much smaller or more distant objects can be distinguished, increased antenna directivity is required. This, in turn, dictates large physical installations, which are difficult and costly to build. To overcome this problem, a new kind of a receiving system, the interferometer,

was developed. This system uses a pair of antennas and transmission lines separated by a specific horizontal distance (i.e., the "baseline") extended to an even multiple of the operating wavelength, preferably more than fifteen wavelengths. The system is usually configured in an east-west orientation. The idea is that a radio signal from a celestial source arrives at two antennas successively, in phase and out of phase, because of Doppler shifts caused by the Earth's rotation. If the two signals are combined through a zero-degree RF combiner, a fringe pattern of interference results, breaking up the large central antenna lobe into a variety of smaller ones. The longer the baseline, the narrower the lobe, or "aperture" (see **fig. 4**).

Very-long-baseline interferometry (VLBI) is possible today through observations made simultaneously by radio telescopes thousands of miles apart, with local oscillators and subsequently recorded data synchronized within a fraction of a microsecond through the use of atomic clocks. This eliminates the need for running coaxial cables from the antenna sites to the central location for processing, and the result can be a beamwidth of 0.0001 arc-second, which is far superior to optical telescopes previously used. For comparison, the 200-inch optical telescope on Palomar Mountain has a theoretical resolution of 0.023 arc-second. Yet because of the effects of atmospheric phenomena, its practical resolution is only about 1.0 arc-second. A block diagram of a VLBI system is shown in **fig. 5**.

Using a special hybrid mapping technique and several radio telescopes located in California, Texas, West Virginia, Massachusetts, and West Germany, astronomers have recently made some exciting new discoveries. The first quasar (3C 147) ever observed with this method has been effectively mapped; it is located some seven billion light-years away. The resolution was in the order of 0.01 arc-second — a considerable improvement over the resolution of the Palomar optical telescope, which detects 3C 147 as no more than a faint star. The radio picture revealed a jet 5000 light-years long emanating from a bright core. Another quasar (3C 273) was observed with a resolution of 0.001 arc-second; its observation recorded matter being ejected from a bright core traveling at nearly the speed of light, a previously suspected phenomenon called *superluminal motion*. (Superluminal motion has been found in two additional quasars and in a distant galaxy as well.)

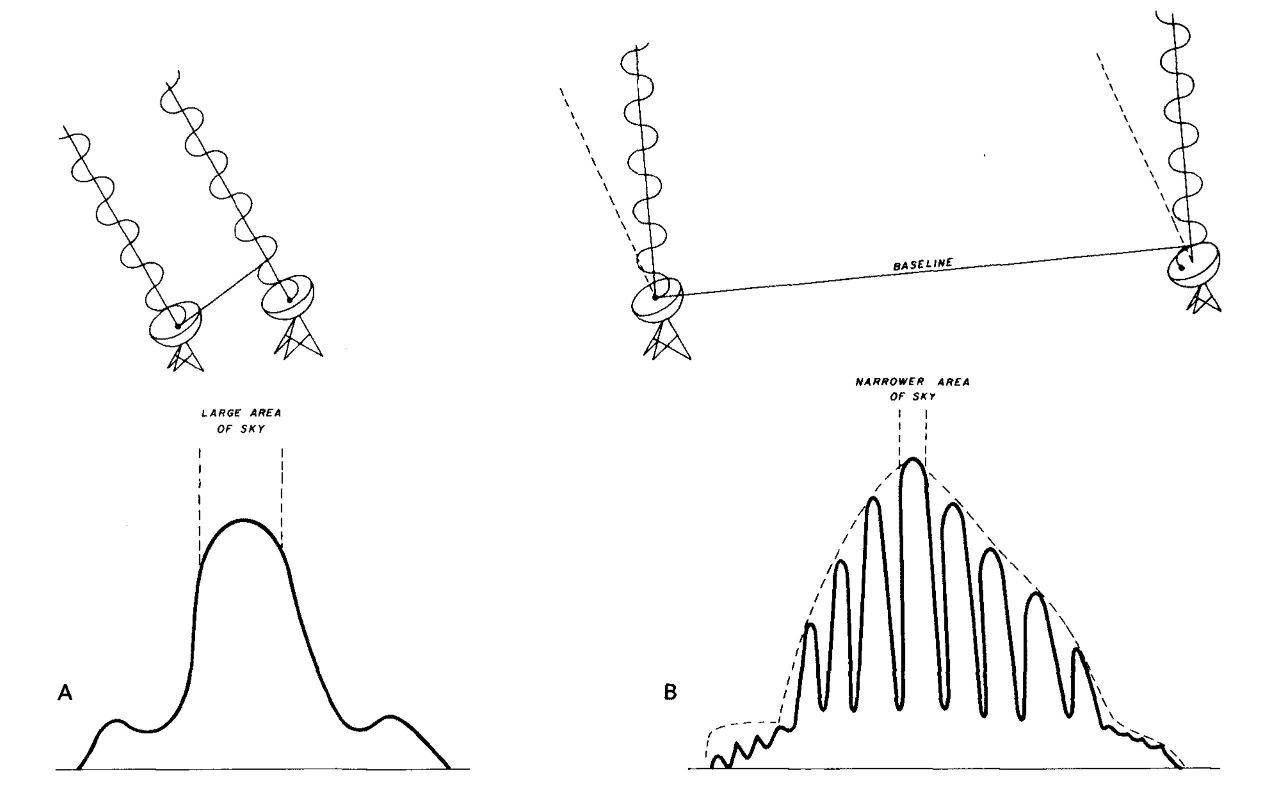


fig. 4. The principle of the radio interferometer telescope. At (A), the signal received by an antenna system peaks over a broad area of the sky. At (B), the signal received by interfering signals from each antenna peaks over a narrower area of the sky.

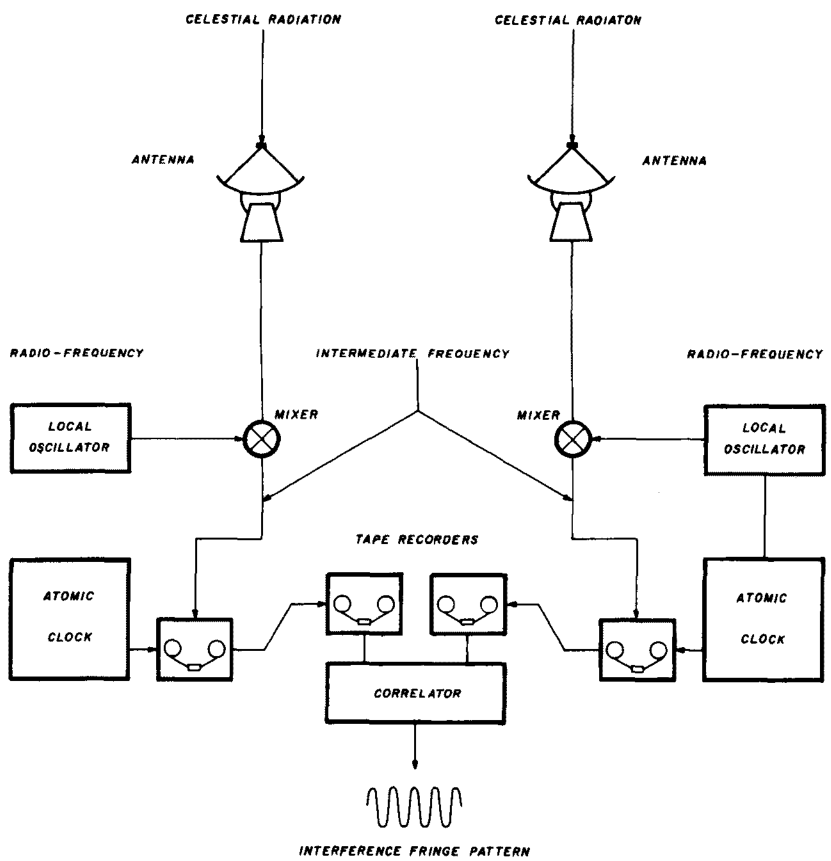


fig. 5. Very-long-baseline interferometry (VLBI) eliminates the need for the antennas to be (hard) connected to a single receiver. Atomic clocks allow the synchronization of separate local oscillators at all locations to a fraction of a microsecond. The final correlation of the data is achieved later at a central processor by playing back and synchronizing the wide-band video information.

We will now turn not to the probability of life elsewhere — the subject has already been discussed in great detail — but to the possibility of receiving intelligent transmissions (see fig. 7). Assume that extraterrestrial life exists. Assume also that the only reason we haven't yet discovered extraterrestrial life is because radio technology has only recently matured sufficiently to allow low noise amplifiers and high resolution microwave synthesizers to be used.

To make the best possible effort in searching for extraterrestrial signals, we would have to cover the microwave frequency range between 10^9 to 10^{10} Hz (1-10 GHz) in narrow steps of, say, 1 Hz. (Ultra-narrow bandwidths are necessary to obtain the best signal-to-noise ratios in SETI.) This means 9×10^9 steps. If we would spend one second per frequency step, and search with one thousand frequencies at once, thereby reducing the number of steps to 9×10^6, it would take approximately three months to search the sky in a single direction. The other condition for our nearly ideal search would be very narrow antenna beamwidths. If our radio telescope would allow a resolution of three million different directions, an all-sky, all-frequency search within the above parameters would be completed in approxi-

Although it is impossible for a Radio Amateur to perform such high resolution experiments, simple backyard interferometers can provide beamwidths as narrow as five degrees, depending upon the frequency of operation and the length of the baseline. (See fig 6.)

the search for intelligence

So far we have discussed only one aspect of radio astronomy, that of receiving and studying wide-band radio noise from non-intelligent sources located hundreds and even thousands of light-years away. The powerful nuclear reactions within these systems result in natural transmissions of formidable amounts of RF power to be detected by our rather primitive radio equipment.

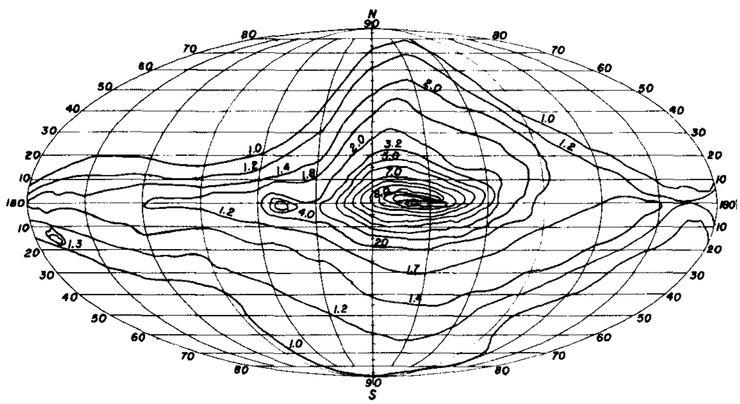


fig. 6. Higher (but still low) resolution mapping of the radio sky than in the case of simple radiometry can be obtained with narrower beamwidth antennas as in the case of amateur interferometry.

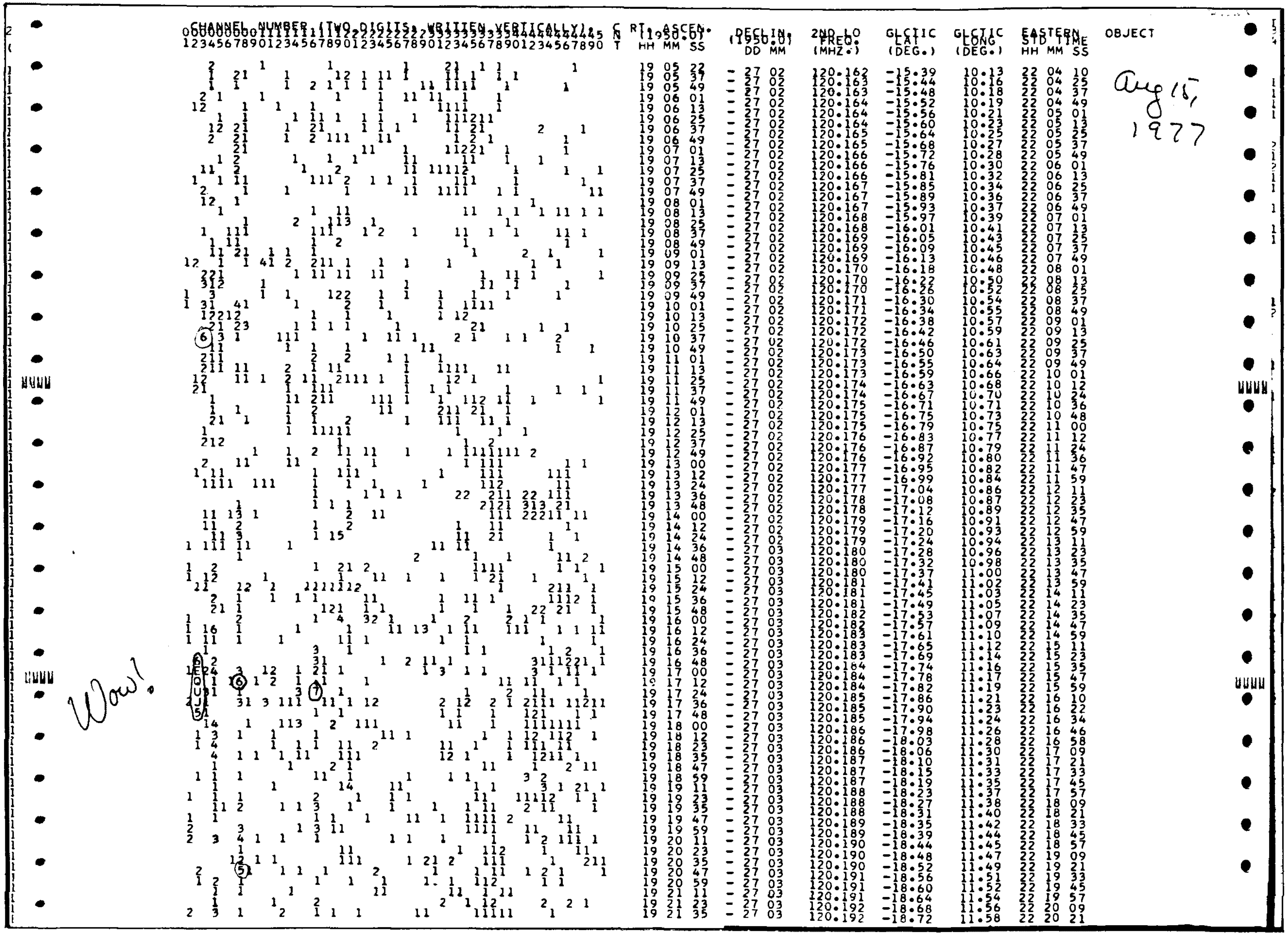
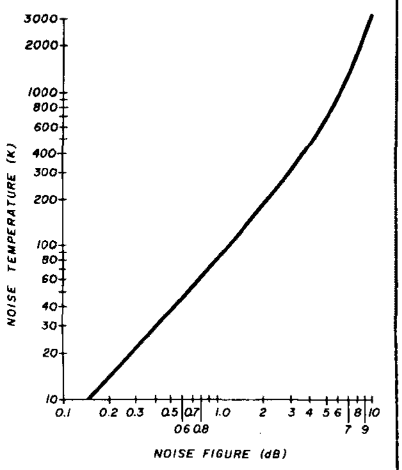


fig. 7. The famous "WOW" signal was received by the Ohio State Radio Observatory on August 15, 1977, at 22 hours 15 minutes and 35 seconds Eastern Standard Time. This signal is one of several mysterious signals received from outer space. Although it lasted only a minute, never to reappear, scientists are certain that the signal was of intelligent origin, and was issued from a source at least as distant as the moon. (*Courtesy of Ohio State University Radio Observatory.*)



$$T = 290\left[10^{\frac{NF}{10}} - 1\right]$$

$$NF = 10 \log\left(1 + \frac{T}{290}\right)$$

fig. 8. The logarithmic relationship between sky noise temperature expressed in degrees Kelvin and equivalent system noise figure expressed in decibels (dB).

mately ten million years — a *rather* impractical proposition for any mortal. We have to reduce the scope of our search, therefore, to match our time and technological limitations. Let's look at some of these limitations.

We can expect that in comparison with the natural RF sources previously discussed, intelligent signals transmitted from outer space would be of much lower power levels. Low power, in this case, could mean extraterrestrial transmitters of powers comparable to our strongest transmitters — one Gigawatt EIRP or more. Consequently, a terrestrial system intended for receiving these signals would have to operate against a quiet background so that its range would be limited only by its own noise figure, which should be no greater than the intergalactic noise level present at its antenna. See fig. **8**.)

Although "intelligent" transmitters could be expected at almost any microwave frequency, radio astronomers have found a quiet range in the frequency *spectrum* that would be ideal for communication with civilizations attempting to communicate with us by radio. (This judgment is based upon our limited idea of what life is. It does not extend to other possibilities such as life forms based on elements other than carbon.)

Located between 1.4 GHz and 1.7 GHz, this area of the spectrum is the "water hole" frequency range. It exhibits a noise temperature of 6 to 8 degrees K (3 to 5 degrees K measured in space). This temperature would allow a 1-Gigawatt EIRP transmitter located approximately 26 light-years away to be heard with a modest backyard SETI radio telescope. (See **fig. 9.**)

The term "water hole" was suggested by the existence of two natural frequencies at each end of the band. Interest in persuing this concept was triggered in 1959 with the publication, in *Nature*, of a paper by Guiseppe Cocconi and Philip Morrison entitled "Searching for Interstellar Communications." Cocconi and Morrison pointed out the importance of radiation from hydrogen atoms reaching the Earth at an ideal spot on the frequency spectrum which coincides with the minimum background noise. At 1.42 GHz there is a natural radio beacon caused by interstellar hydrogen (H); another natural beacon exists at 1.66 GHz. This one is caused by hydroxyl (OH) ions traveling in space. When chemically combined on Earth, the two produce water (H_2O) — thus the terminology "water hole." Because hydrogen is the simplest, most abundant element in the universe, and because water is one

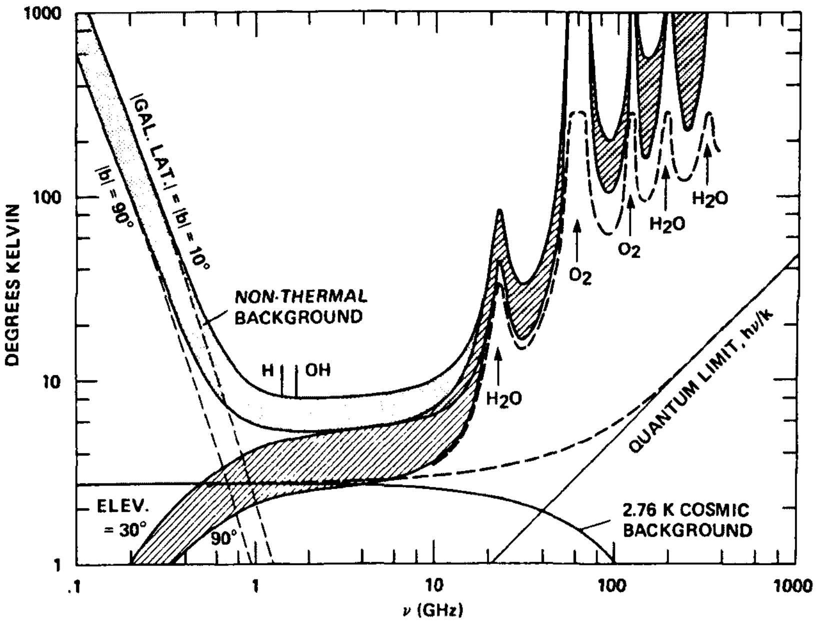


fig. 9. The "water hole" window provides the best overall noise performance for receiving intelligent transmissions from outer space. This frequency band has been suggested by astronomers as a possible radio meeting point for extraterrestrial civilizations which base their existence on water (see text).

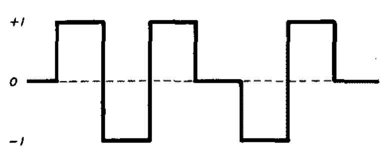


fig. 10. The concept of information obtained at the output of a circularly polarized, sense-switched receiver which is responding to a hypothetical binary circular polarization modulated signal and using a single frequency.

of the basic requirements of life as we know it, this frequency range has been favored by scientists as the "magic" band for interstellar communications. The concept of the water hole assumes two things: first, that all life in the universe is a function of water, and second, that any extraterrestrial civilization attempting to communicate with us would select this frequency band for the same reasons we did.

One important factor in receiving intelligent transmissions would be the signaling protocol and the rate of transmission used by the sending civilization and consequently the modulation scheme. If we may judge by our own experience, it is reasonable to assume that the sender would choose a simple two-state binary signaling scheme that could be modulated slowly (and therefore compatible with signal-to-noise bandwidth requirements) in one of four modulation schemes: amplitude, frequency, phase, and polarization.

A careful analysis of these modulation techniques indicates that the first three would be difficult to receive. If amplitude modulation were used, a binary "1" would be detected as the transmitter would be turned on. However, positive identification of the reverse state (i.e., 0) would be less probable because there would be no signal to reveal this information. While this method is acceptable in casual CW signaling, anti-cryptographic studies indicate that information would be lost if such a method were used (a true −1 state would be required for positive identification).

Two distinct binary states could be obtained with conventional frequency shift keying. However, the introduction of a new element — the second frequency — would make the search more difficult in view of the narrow bandwidths used. While phase modulation is a superior method for carrying data communication in that it requires only half the signal-to-noise ratio of the other modulation schemes for the same amount of information, it is thought to be the least likely to be used in searching for unknown signals.

The most likely method of radio communication that might be used by an extraterrestrial civilization is binary antenna polarization modulation using the same frequency. By properly changing between two orthogonal polarizations such as two perpendicular linear polarizations, or between left and right circular polarizations, the two binary states could be transmitted on the same frequency by switching the transmitter's output as shown in fig. **10**. This in turn would allow for reversely polarized receiving antenna arrays on earth to receive the binary information and process it through two distinctive radio receivers as shown in fig. **11** — or one receiver that would switch between two properly polarized antennas. Most searches for intelligent signals to date have been performed in the water hole frequency range using the latter method.

designing receivers

Over the past few years several methods have been suggested for receiving ETI signals. One technique — based on the "pulse" theory — stands a good chance of acceptance and is of interest to Amateurs because it requires simpler receiving equipment than other methods. This technique assumes the transmission of high power pulses of one second or longer in a digital binary format, as previously discussed. This concept makes sense because the average power available from a hypothetical extraterrestrial transmitter would probably be limited by thermal inefficiencies. (Although the topic is debatable, we assume that extraterrestrials would have technological problems similar to ours.) Much more peak power could be obtained from pulsed binary transmitters, which can overcome the noise figure limitations of target receivers and can be spread over relatively wider bandwidths so that complicated Doppler corrections would be minimized. Pulse receivers with ultimate bandwidths of up to 10 kHz have been used in the "magic" frequency range.

On the other hand, recent experiments favor the very narrow bandwidth/beamwidth beacon approach because of the superior signal-to-noise ratio obtainable. Using this concept, powerful beacons would be directed at the solar system chosen as an appropriate "target" by the sending civilization. The signal would be transmitted frequency-corrected so that it would be received on earth near the laboratory neutral hydrogen-line frequency (1.42 GHz/21 cm), thereby simplifying our search. The correction would include the source's Doppler shift and the frequency shift caused by the radial velocity of our sun, known by the sender from long-term astronomical observations of our solar system.

A terrestrial receiver would be equipped with multiple ultra-narrowband IF filters. The theoretical minimum bandwidth for interstellar communications has been recently calculated by Drake and Helou[2,3], who indicate that its limit is determined by the effects of the multipath scattering phenomenon resulting from turbulent ionized gases — the same effect that causes pulsar scintillation. Consequently, a pure carrier in the water hole will have a tendency to be wider at the arrival point than its originated bandwidth, say 0.01 Hz, after traveling 100 Parsecs (1 Parsec = 3.26 light-years). This can be important in the selection of bandwidth set as the limitation for practical filter design in SETI receivers.

Ultra-narrowband searches have been conducted at the Arecibo Observatory and at Harvard University by a ham radio operator, Professor Paul Horowitz, W1HFA. In 1978 Horowitz observed nearly 200 nearby stars similar to our sun at the 21-cm neutral hydrogen line wavelength by using the observatory's 305-meter dish (see fig. **12**). The receiver (figs. **13A**, **B**, and **C**) used a dual circular polarization feed which gave an effective antenna area of 22,000-m^2 (with 68-dB gain). The overall system temperature was 80 degrees K.

As shown in fig. **13D** parametric amplifiers were used for the front end (new GaAs FET amplifiers have since

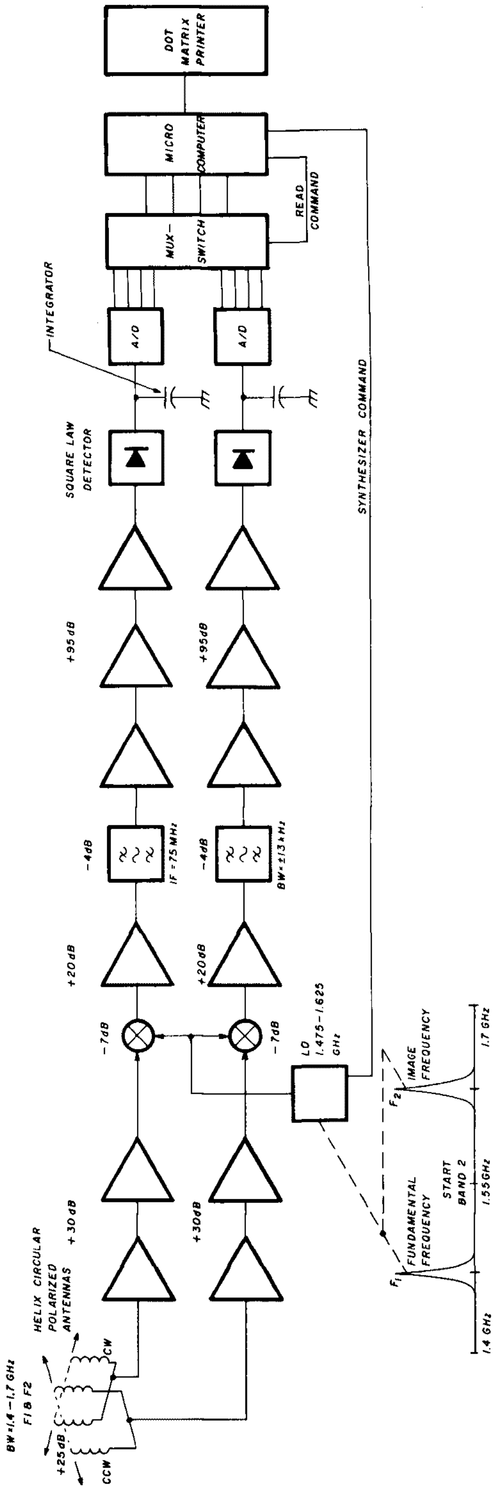


fig. 11. Block diagram of the author's SETI receiver, under construction. The system assumes circular polarization modulation (clockwise and counterclockwise). Dual diversity is achieved through the strategic choice of the 75-MHz IF, which allows image recovery in both modulation senses. Consequently, the entire 300-MHz water hole frequency band can be viewed in half the time it would otherwise take. Final selectivity is achieved with two crystal filters (4171-F monolithic from Piezo Technology) at 75 MHz. The bandwidth is ±13 kHz. The signal-to-noise problem introduced by the image recovery technique can be compensated for through longer integration time. The preamplifiers have a calculated noise figure of 0.4 dB (NEC NE 71083 and NEC 72084).

been installed) along with a dual conversion scheme having a first IF at 260 MHz *and a second IF at 30* MHz. Crystal filters with bandwidths of 5 kHz were present in the second IF, followed by IF amplifiers. The final conversion to the sine and cosine basebands was achieved with a 30-MHz third oscillator. All local oscillators in the system were synthesized with a final resolution of 0.01 Hz and a short-term stability of $\Delta f/f = 5 \times 10^{-12}$ provided by a rubidium-referenced clock.

The quadrature baseband signals were then filtered by four-pole Butterworth low-pass tunable filters which were sampled under computer control with an *analog multiplexer* and 12-bit analog-to-digital converters. A single observation consisted of 64-K (65,536) complex samples at 1 millisecond intervals from each of two polarizations. The samples were digitized and recorded in real time onto nine-track digital magnetic tape for follow-up processing, which included a fast Fourier transform (FFT) on the quadrature signals.* A complete 64-K complex FFT and computation of power spectrum with the instantaneous resolution of 0.015-Hz per bin (compatible with minimum theoretical bandwidth discussed earlier) took about 15 seconds.

In the observations, it was assumed that a true celestial signal would have already been frequency-compensated for Doppler by the originating civilization. This included the sender's own velocity and the radial velocity of our solar system as observed by the sender, but not the residual Doppler resulting from the Earth's own rotation (not known to sender), which can shift such a signal through another 0.15 Hz of bandwidth at 1.5 GHz. This correction was implemented in the receiver's synthesized local oscillators (first LO and third LO) which were swept via real-time computer control — that is, the local oscillator was updated several thousand times during each observation. The synthesizer control mechanism set the first LO frequency at the beginning of each observation so that the third LO began at 30 MHz; the third LO was then updated at 20-millisecond intervals by computing frequency offsets in real time according to a polynomial algorithm which approximated the Earth's velocity according to data obtained from the Lincoln Laboratory planetary ephemeris.

One positive side effect resulting from this frequency sweeping through the ultranarrow bandwidths of the receiver was that earth-generated interference (which is *generally not* frequency swept) is completely rejected by the system. This was confirmed throughout the search by the absence of false alarms.

The Arecibo experiments revealed no evidence of ETI: Today the search

*FFT is usually used to break down complex waves.

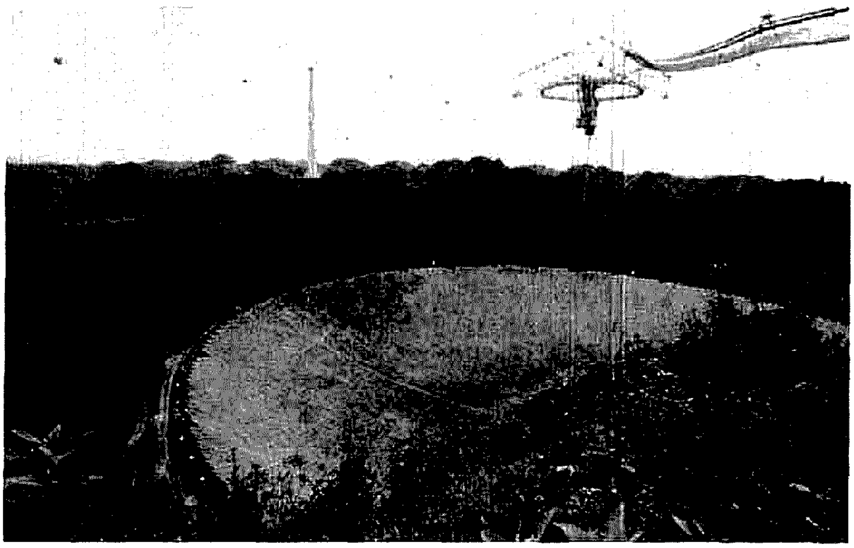


fig. 12A. Overview of the Arecibo Observatory, where in 1978, Paul Horowitz, W1HFA, observed nearly 200 nearby stars similar to our own sun at the 21-cm neutral hydrogen line wavelength. ***(Photo courtesy of College of St. Thomas, St. Paul, Minnesota.)***

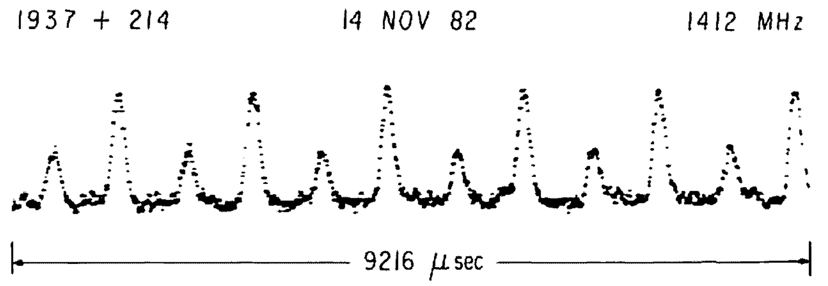


fig. 12B. Pulsar profile PSR 1937+214, discovered at Arecibo in 1982, measures only 3 miles across but releases 10 to 100 million times the energy of the Sun. Rotating around its gravitational axis at 642 revolutions per second, it pulses at 1.558-millisecond intervals; each pulse consists of two signal peaks or flashes of radio energy streaming out ***from its magnetic poles.*** **Pulsars** ***were*** **first thought to be of intelligent origin because of their precise repetition rate. At Cambridge University in 1967, a signal was recorded at 1.33730-second intervals. Code-named** ***LGM*** **— "Little Green Men" — it was later proved to be the first known pulsating neutron star, or pulsar.** ***(Courtesy of Arecibo Observatory, part of the National Astronomy and Ionospheric Center operated by Cornell University under contract with the National Science Foundation.)***

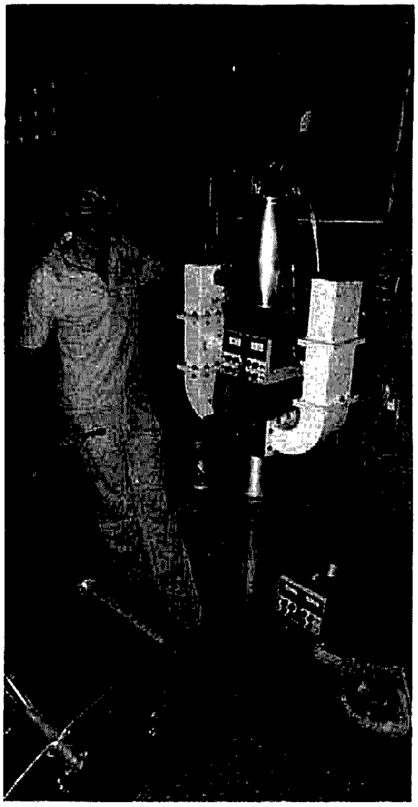


fig. 13A. Inside the carriage house of the Arecibo Observatory — suspended above the 1000-foot dish — engineer Bob Zimmerman, NP4B, proudly displays the new dual-channel 18-cm receiver front end (left and right circular polarizations). (*Photo courtesy of Arecibo Observatory.*)

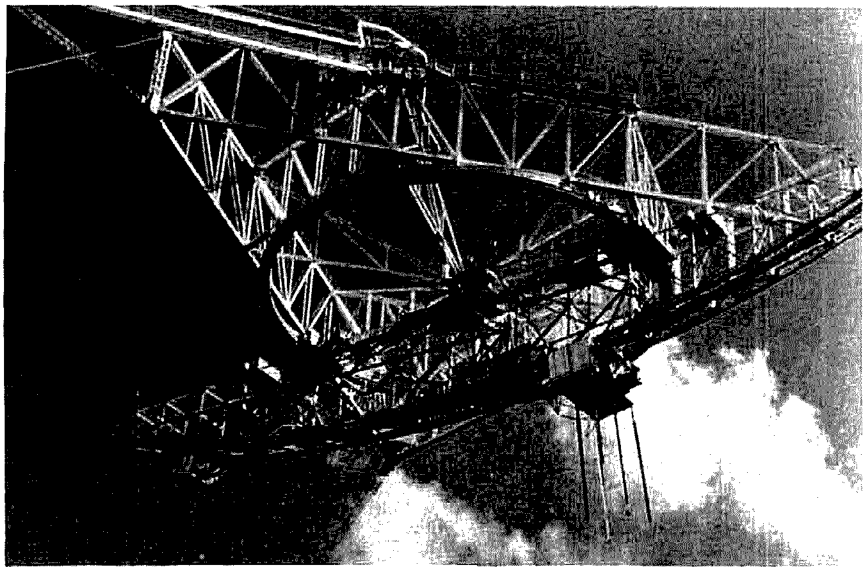


fig. 13B. The antenna feed passes through the floor of the carriage house and is focused in the dish, 430 feet (131.06 meters) below. The temperature inside the metal dome is held at 80 degrees Kelvin with liquid nitrogen in order to reduce the noise temperature of the front end. (*Photo courtesy of College of St. Thomas, St. Paul, Minnesota.*)

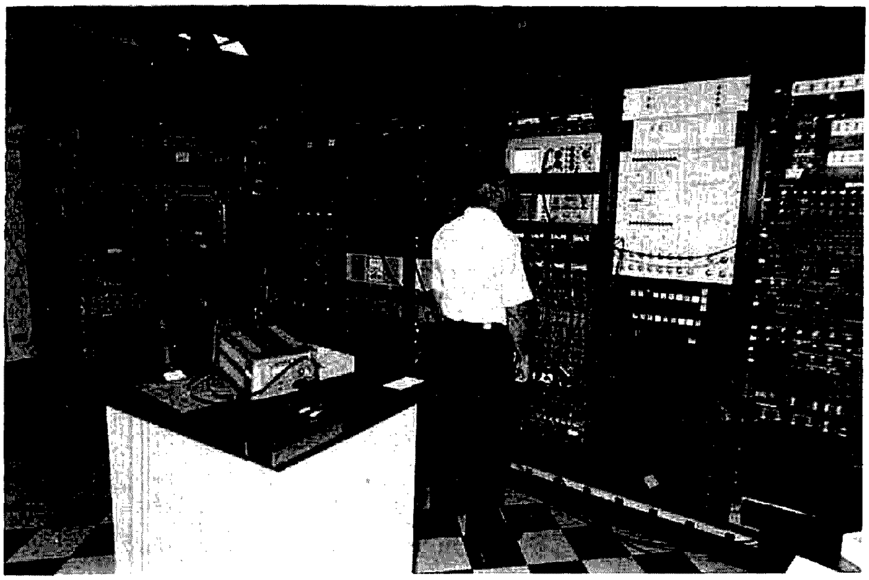


fig. 13C. The receiver room at Arecibo Observatory. The equipment contains oscillators, HP synthesizers, Rubidium standards, detectors, amplifiers, A/D converters — just about any electronic device needed by an astrophysicist. About 200 hours per year are dedicated to SETI. Searches have been made of approximately 1000 nearby stars at the Water Hole frequency. (*Photo courtesy of Arecibo Observatory.*)

continues with a special receiver in operation at the Planetary Society/SAI/Harvard project "Sentinel" as shown in **fig. 14**. This system matches the natural minimum bandwidth discussed earlier by also resolving the input bandwidth into 64K (65,536) complex frequency bins of 0.03 Hz each. The 84-foot radiotelescope is equipped with two dual-circularly polarized feedhorns (5 bands) connected to two receivers. The front end consists of two identical 35-dB gain, 55 degrees K (uncooled), 10 degrees K (cooled) GaAs FET preamplifiers operating in the waterhole frequency band. (Other frequencies can be tuned.)

The receiver uses a conventional single conversion scheme with an IF of 30 MHz. Image rejection mixers, broadband IF amplifiers and filters are used in conjunction with a computer-controlled synthesizer. The 30 MHz IF signals are then sent to the back end of the receiver, located in the control building via low-loss rigid coaxial cables. As in the case of Arecibo, the back end is responsible for sweeping through the 30 MHz IF to compensate for the Doppler shift caused by the Earth's rotation. The result is a quadrature baseband combination of

signals which is further filtered through 6-pole low-pass anti-aliasing filters. The control computer updates the LO 40 times per second based on an ephemeris table calculated at the beginning of each run. Sample-and-hold amplifiers and 8-bit analog-to-digital (A/D) converters are used to feed the FFT processors via interrupt-driven parallel ports.

Although the Harvard installation surpasses, by at least an order of magnitude, the combined efforts of all previous SETI (in terms of system sensitivity, the number of sky positions observed, and the number of concurrent channels), scientists feel that the search should be expanded in frequency by a factor of at least 100.

This would mean increasing the present 64-K channels used in each of two polarizations to about 8.4 million channels of 0.05-Hz resolution, thus increasing the probability of intercept (POI) by a factor of 100. Consequently, the instantaneous bandwidth would increase from the present 2 kHz to 420 kHz. Although this would be quite an improvement, it would still be insufficient to cover the 300-MHz bandwidth of the water hole at once. An all-sky water hole search with the new receiver would still require 1200 instantaneous bandwidths of 420 kHz times the number of sky locations.

Because no receiving system can cover the entire sky at all frequencies at once, much more work remains to be done in SETI, and while omnidirectional wideband pulses have been suggested as a SETI method, the narrowband beacon concept gives superior S/N ratios not achievable otherwise. On the other hand, Doppler corrections associated with the beacon approach, which would require hard-to-design high-resolution microwave synthesizers, make the pulse concept attractive at least for the Radio Amateur. New methods of observing many RF sources simultaneously using Bragg-Cell technology have been suggested. However, the relatively wide channel bandwidth produced by today's Bragg technology, combined with the low receiver dynamic range, limits the applicability of this technology.

We have looked at several radio astronomy systems, from a simple radiometer to the ultra-narrowband receivers used by professional radio astronomers. Although this article is not intended as a construction paper (ample details are provided in the references), some elements of design should be considered before a system approach is chosen. The block diagram shown in **fig. 15** shows an economical approach to designing an Amateur Radio astronomy center operating in the water hole frequency band. It could be used as a wideband radiometer, an interferometer, or as a tunable narrowband receiver intended for the reception of pulses if care is taken in providing short-term stability for the local oscillators along with narrowband filtering in a third conversion.

This system would consist of a two-stage GaAs FET preamplifier with a low noise figure. Several designs have been recently published in the literature. The expected gain from such amplifiers is typically in the 30 dB range or better. This would be sufficient to overcome the high noise figure of the following mixer (7 dB). Recent designs using the Mitsubishi MGF 1412-11-09 and MGF-1412-11-10 GaAs FET transistors claim noise figures of about 0.5 dB (35 degrees K). Older designs pro-

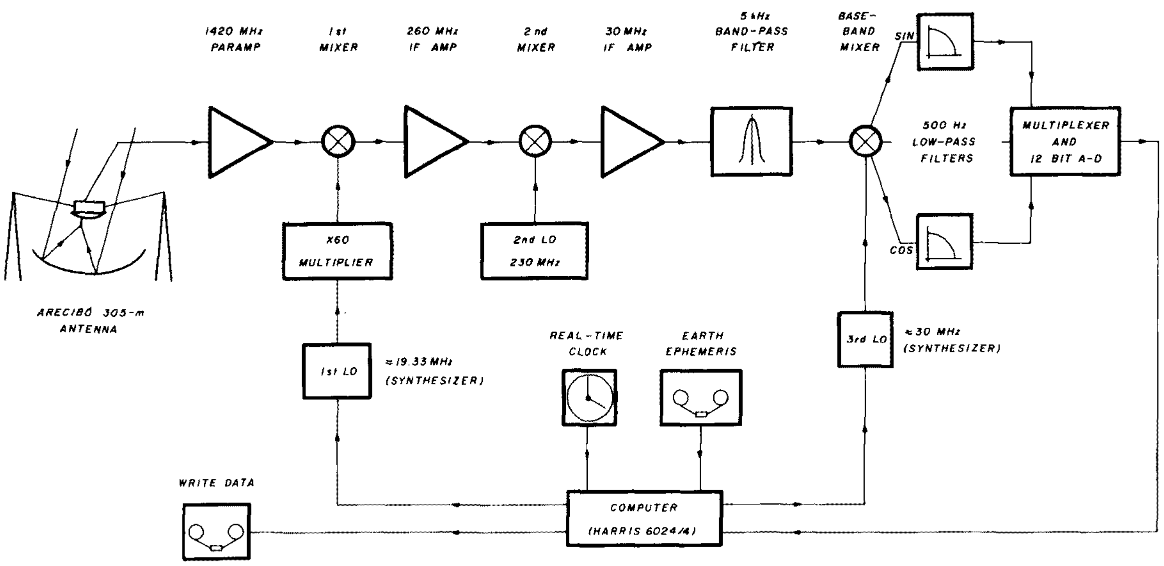


fig. 13D. Block diagram of the Arecibo system as used for SETI in 1978 by Professor Paul Horowitz (W1HFA). Frequency sweeping of the local oscillators was used to compensate for Doppler shifts caused by the earth's rotation. (See text).

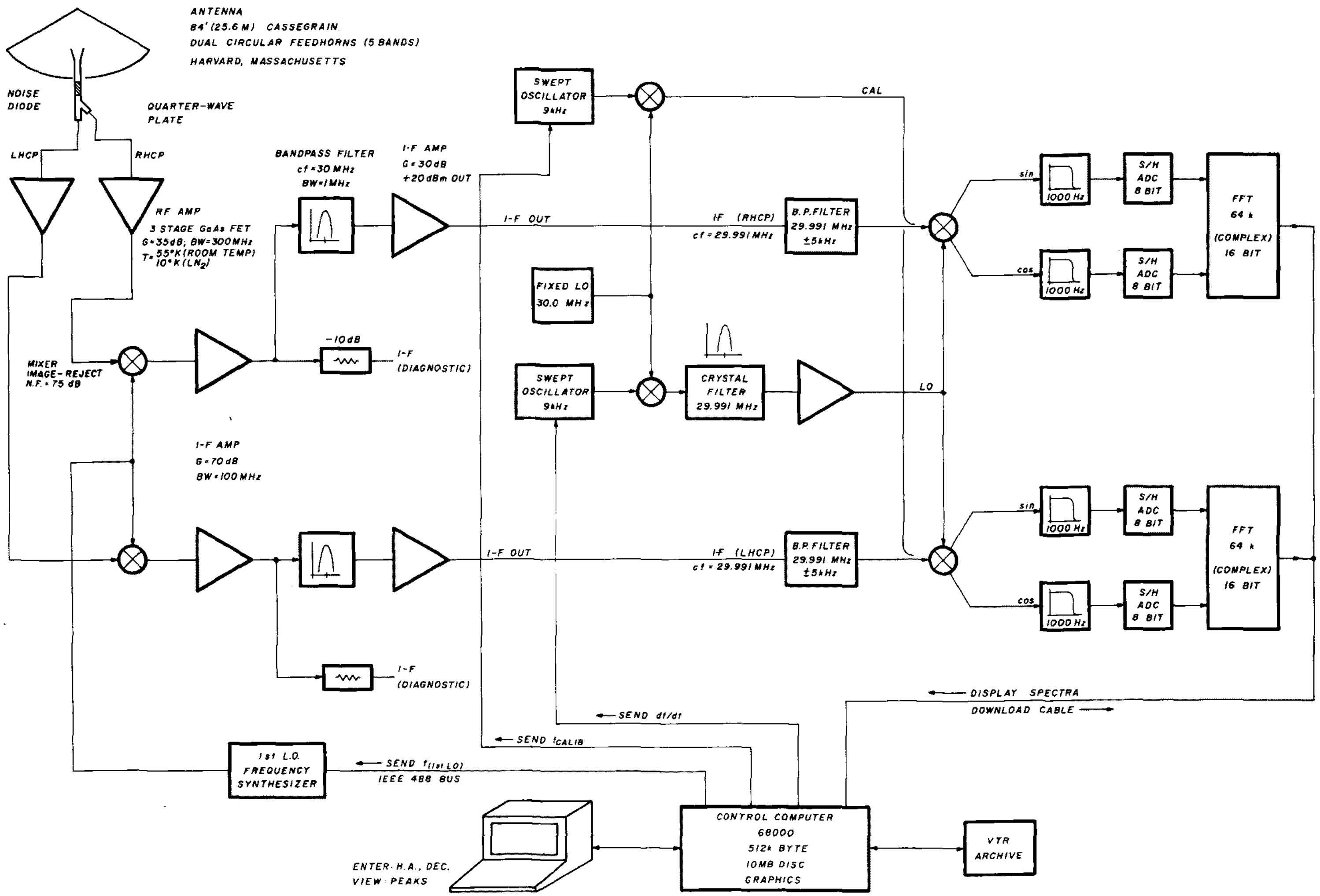


fig. 14. Block diagram of the Ultra-Narrowband SETI receiver used at Harvard University. The system uses an 84-foot (25.6 meter) equatorial radiotelescope with a single conversion dual polarization receiver intended for processing left and right circular polarizations at the same time. (*Courtesy of Harvard University Physics Department.*)

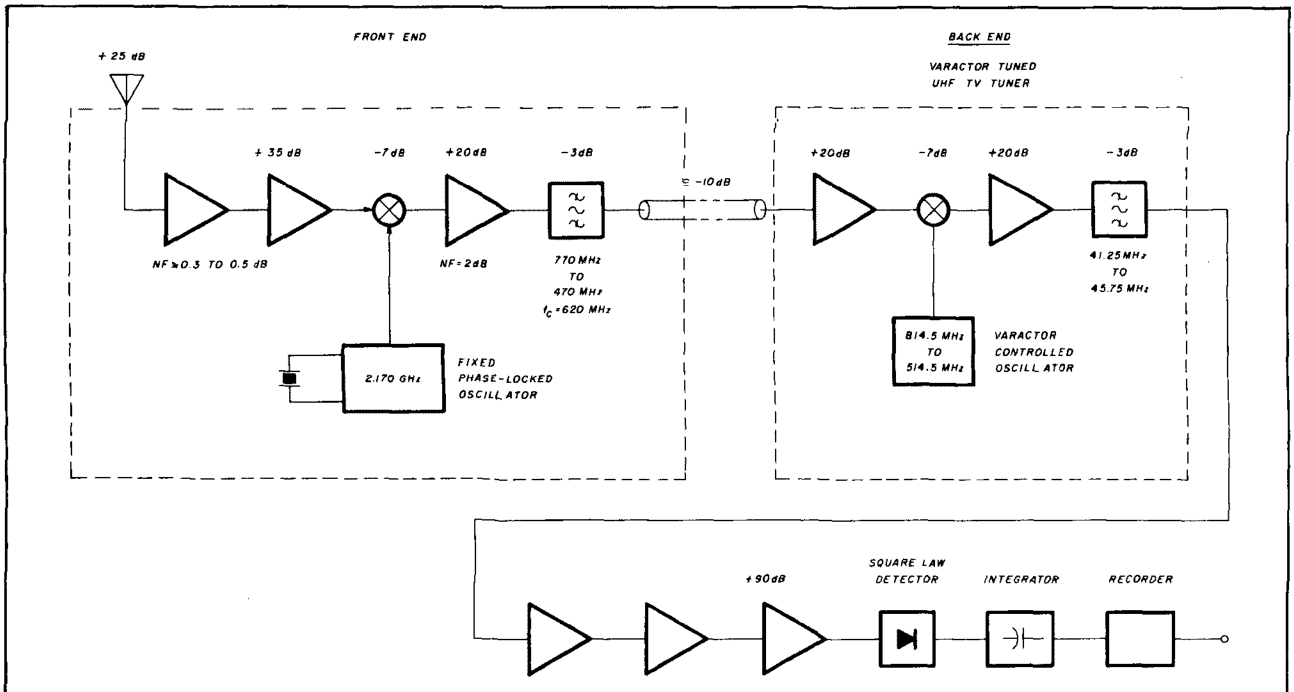


fig. 15. Block diagram of an economical amateur radio astronomy receiver.

vide noise figures in the 0.8 dB (58 degrees K) range with about the same gain. Stripline approaches on G-10 material or open frame designs have been extensively covered in the reference material at the end of this article along with simple rat-race mixers intended for 1.296 GHz or 2.3 GHz. (Mixers can also be purchased from a variety of manufacturers.)

Our receiver would use the first mixer in conjunction with a fixed phased-locked loop or multiple chain synthesizer at 2.170 GHz. (Such designs are popular in satellite converters and other devices.) The first IF would be purposely chosen to fall in the UHF range where an inexpensive varactor-type TV-tuner could be used as a second tunable IF over the entire water hole frequency of 300 MHz. The part investigated was a Mitsumi UES-A 56F, which is marketed by several companies, including Radio Shack. This tunable converter exhibited a total noise figure of about 6 dB and 15 dB of gain, with an acceptable short-term stability at room temperature compatible with an ultimate bandwidth of 10 kHz (care should be exercised to provide isolation of this unit from the front end because harmonics of local oscillator fall in the input range of the receiver).The tests were performed with a precision power supply having a range of 0 to 28 volts. If only a radiometer is contemplated, the phase-locked first LO is not mandatory and the control voltage applied to the second converter should be chosen about halfway on the voltage curve and should be double regulated.

The UHF converter provides a TV-compatible IF, centered at 44.5 MHz, with a bandwidth of 6.5 MHz. This output could be used directly with a modified high-gain TV-IF with the

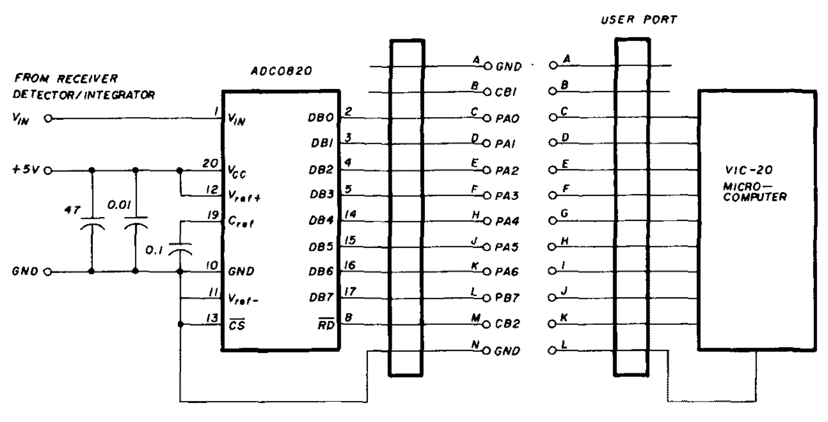


fig. 16. Schematic diagram of a simple A/D interface for a radio telescope. No additional parts are needed to perform consecutive readings; the information can be printed out as relative flux data by a dot-matrix printer directly from the microcomputer.

the candidate stars:
which stars might support life?

Although our knowledge is limited, it appears that the universe is expanding — that is, the galaxies are moving apart from each other. This movement suggests a "time of beginning" in a cycle (known as the "big bang" event) that began with an explosive fireball of matter inside a huge black hole with no conceivable limits some fifteen billion years ago.

Certain stars of various sizes evolved from the cold gas of a previous cycle; these are now in their "main sequence," but approaching a "finale" as shown in **fig. A.** Depending on their mass and temperatures, they are classified by letters, with the hottest designated by the letter O, and followed in descending order by other spectral types such as B, A, F, G, K, and M. In the search for extraterrestrial intelligence only type F, G, and K stars are of interest to us because they are the right size and temperature for supporting life on planets similar to our own. (Our sun is a type G2 yellow star.) With some three hundred billion stars in the Milky Way galaxy alone — and ten billion other galaxies in the known universe — we can identify approximately one million nearby candidate stars (within 1000 light-years) of spectral type F, G, or K that could conceivably support life. Despite the magnitude of this number, only a handful of stars (see **table 1**) are within the Amateur's technological reach. Of this handful, the nearest are in Alpha Centauri. Located some 4.3 light-years away, Alpha Centauri is a triple system containing two massive suns (Type G4 and Type K1) separated by some 20 astronomical units (1 A.U. equals the mean distance of the Earth from the sun) and revolving around each other along with a smaller third star, Alpha Centauri C (a type M star). Recent investigations indicate that this system may be much younger than ours, suggesting that advanced forms of life would probably not have developed even if a planetary system did exist within its complex rotational setup. Other theories, however, might explain a heightened probability of life on Alpha Centauri as shown in **fig. B.**

Of approximately 40 stars located not more than 16.7 light-years from Earth, only two — Epsilon Eridani (type K2) and Tau Ceti (type G4) — have been identified as meeting the conditions necessary for the existence of advanced forms of life. Similar to our sun but somewhat smaller, Epsilon Eridani is located 10.5 light-years away in the constellation Eridanus. Tau Ceti is located approximately 10.8 light-years away in the constellation of the Whale. Rather dim compared to our sun, it is visible from Earth's northern hemisphere only during the winter months. These two stars were among the first to be observed by Frank Drake and his team at the National Radio Astronomy Observatory (NRAO) in Green Bank, West Virginia, in 1960 under the project name *Ozma*, a name borrowed from the Wizard of Oz. No evidence of extraterrestrial intelligence has been observed to date.

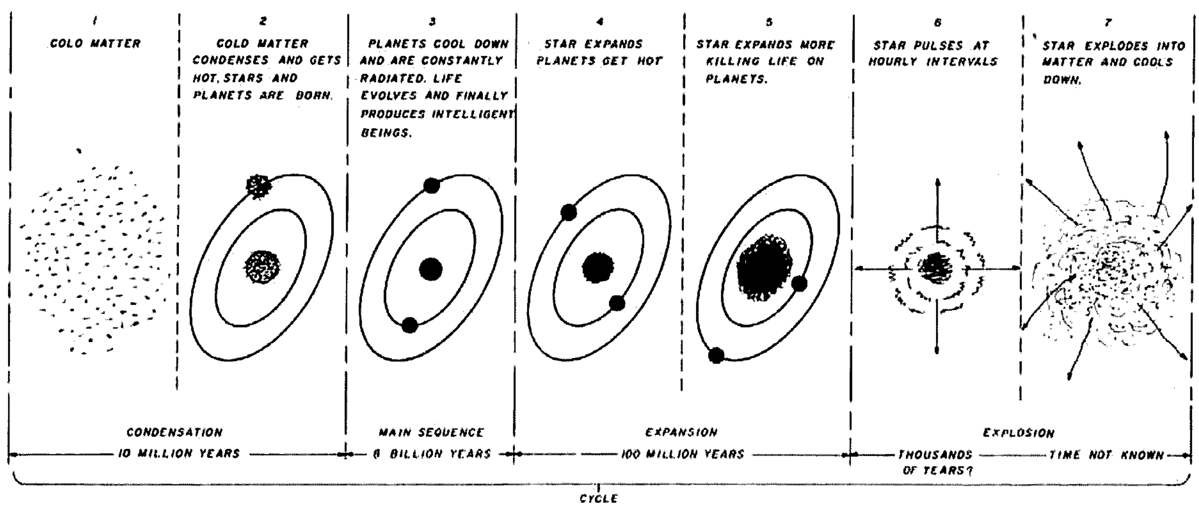


fig. A. One concept of star evolution suggests that the possibility of life on planets revolving around a type G yellow star occurs during its "main sequence". This type of star is similar to our own sun. A "lucky" planet similar to Earth would need approximately four billion years of continuous energy flow from this star to allow the random process of mutation on elements to produce the complexity of the human brain, which has made possible the development of communication technologies.

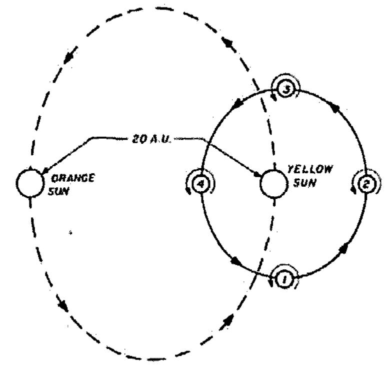


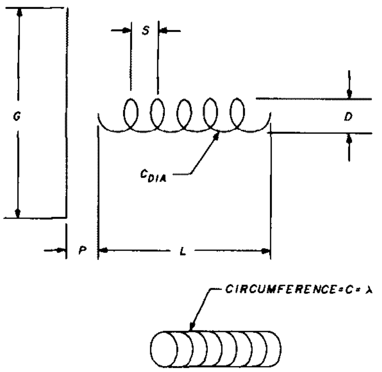
fig. B. The concept of two suns — one yellow (Alpha Centauri A, type G) and the other orange (Alpha Centauri B, type K) — rotating around each other in the triple star system Alpha Centauri, located some 4.3 light-years away. If an earth-like planet were to exist at the same distance from the yellow star as the earth is from the sun, it is conceivable that the complex rotational relationship would allow for long alternating yellow and orange days — with no nights — which may accelerate the development of life on this relatively young system.

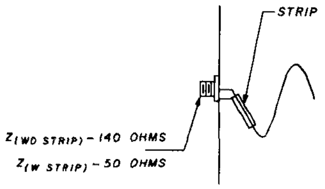
table 1. Stars within 26 light-years which could have habitable planets (adapted from Stephen H. Dole, 1964.)

name of star	spectral class	approximate distance (light years)
Alpha Centauri A	G4	4.3
Alpha Centauri B	K1	4.3
Alpha Centauri C	M1	4.3
Lai 21185 (A)	M1	8.2
ε Eridani	K2	10.8
61 Cygni A	K5	11.1
61 Cygni B	K8	11.1
ε Indi	K5	11.3
Grm 34 A	M2	11.7
Lac 9352	M1	12.0
τ Ceti	G8	12.2
Lac 8760	M0	12.6
Cin 3161	M3	14.9
Grm 1618	K8	14.9
CC 1290	M3	15.4
Cin 18,2354	M3	16.1
+15° 2620	M1	16.9
70 Ophiuchi A	K1	17.3
70 Ophiuchi B	K5	17.3
η Cassiopeiae A	F9	18.0
η Cassiopeiae B	K6	18.0
σ Draconis	G9	18.2
36 Ophiuchi A	K2	18.2
36 Ophiuchi B	K1	18.2
36 Ophiuchi C	K6	18.2
HR 7703 A	K2	18.6
HR 5568 A	K4	18.8
HR 5568 B	M0	18.8
δ Pavonis	G7	19.2
−21° 1377	M0	19.2
+44° 2051 A	M0	19.2
+4° 4048 (A)	M3	19.4
HD 36395	M1	20.0
+1° 4774	M2	20.2
+53° 1320	K7	20.2
+53° 1321	K9	20.2
−45° 13677	M0	20.6
82 Eridani	G5	20.9
β Hydri	G1	21.3
HR 8832	K3	21.4
+15° 4733	M2	21.8
ρ Eridani A	K2	22.0
ρ Eridani B	K2	22.0
HR 753 A	K3	22.0
Vega	G4	26.0

One of the more recent additions to the list of "interesting" stars is Vega, located 26 light-years away from Earth in the constellation Lyre (Lyra). Vega is the third brightest star in the sky. Although twice the size of the sun, its surface temperature has been measured and found to be almost the same. A relatively young star — at only one billion years old — Vega is important to us because of the discovery, in 1983, of a possible planetary system around it by the infrared astronomical satellite (IRAS). According to astronomers, while the infrared telescope aboard IRAS was sensitive enough to detect a mass rotating around Vega equivalent to the combined mass of all nine planets in our solar system, it could not resolve the objects precisely enough to distinguish among them. Nonetheless, this is one of the most compelling pieces of data suggesting that we may have another planetary system in the universe. (This theory is now being challenged by another interpretation; some investigators view the phenomenon as a belt of dust consisting of "pellets" that reradiate the star's infrared energy.)

sound trap removed or with a home-brewed IF of a lower noise figure. Full rectification could be implemented for detection in conjunction with simple integrators and DC amplifiers. Fast Fourier transforms (FFT) could also be implemented by using inexpensive microcomputers. The receiver's output transducer could be one of the newer A/D converters — such as National's ADC0820 — connected to a microcomputer and a dot matrix printer. This converter eliminates the extra circuitry normally associated with interfacing A/Ds to microcomputers as shown in **fig. 16**. It was specifically designed to appear as memory locations or I/O ports to a standard microprocessor, with no other logic needed. In addition, the converter's input acquisition time is much faster than its conversion time, lending its use to measuring many analog signals upon software commands without the aid of additional sample-and-hold devices. The resolution is 8 bits, with a





$C_{DIA} = 0.017\lambda$	*conductor diameter*
$D = \frac{\lambda}{3}$ OD	*loop diameter*
$G \geq 0.8\lambda$	*ground plane*
$N \geq 3$	*N = 9 for 14.8 dB* *N = 12 for 16 dB*
$S = \frac{\lambda}{4}$	*space between turns*
$\lambda_{(IN)} = \frac{11{,}810}{f(MHz)}$	*wavelength free space*
$P = 0.12\lambda$	*space from reflector to first turn**

**observe impedance matching stub*

(A) Theoretical formula for designing a helical antenna.

F_{RANGE} = 1400 MHz to 1700 MHz
BW = 300 MHz
C_f = 1550 MHz
if C = λ

$$BW\ is \left| \begin{matrix} & 1550\ MHz & \\ -20\% & C_f & +30\% \end{matrix} \right|$$

ANTENNA BANDWIDTH 3 dB POINTS (C_f = 1550 MHz) IS:
BW 1240 MHz to 1860 MHz

$$\lambda = \frac{11{,}810}{1550} = 7.6193\ INCH$$

$$C_{DIA} = 0.017\lambda = 7.619 \times 0.017 = 0.1295\ inch$$

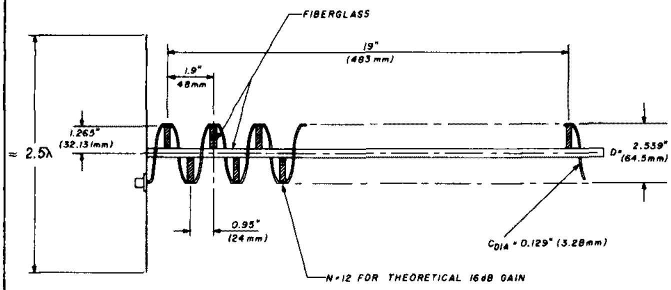
$$D = \frac{\lambda}{3} = \frac{7.6193}{3} = 2.539\ inch$$

$$G \geq 0.8\lambda \geq 0.8 \times 7.6193 \geq 6.09544\ inch$$

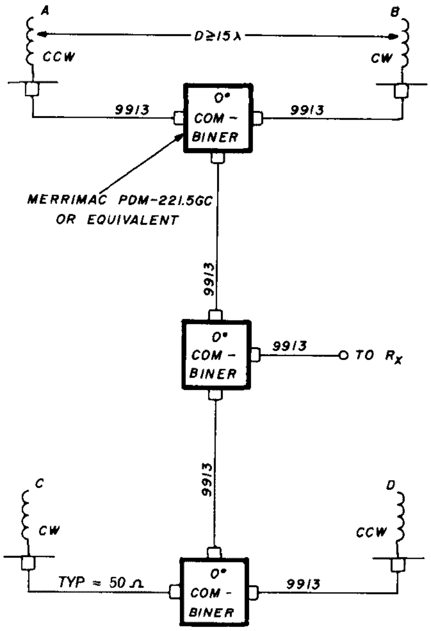
$$P = 0.12\lambda = 0.12 \times 7.6193 = 0.9143\ inch$$

$$S = \frac{\lambda}{4} = \frac{7.6193}{4} = 1.90\ inch$$

(B) Actual design of a helical antenna for the water hole frequency band.



(C) Construction details.



summary of polarization combinations

A + B → Horizontal
A + C → Vertical
A + D → Clockwise
B + C → Counterclockwise
A + B + C + D → Linear

(D) An array of four helical antennas can be arranged to provide narrow beam width and many different polarization patterns. Additional LNA's can be introduced in each of the antenna legs to overcome the noise figure of the combiners.

fig. 17. Design of a helical antenna system for the water hole frequency range.

maximum conversion time of 1.2 microseconds, an ideal application for a radio astronomy center. The ADC0820 chip was tested with the VIC-20 microcomputer, but would be equally applicable to any microcomputer having a latched data bus. With the proper software, the output of the computer can be printed out at equal time intervals in relative flux units on a scale of 0 to 5 volts (or 0 to 9 volts with a modified reference).

what antenna to use

The best antenna for radio astronomy is still the parabolic dish, as proven by most professional radio astronomy centers. Some Amateurs are reportedly using computer-controlled steerable dishes as large as 60 feet in diameter. But unless Amateurs have access to large backyards and friendly neighbors, they cannot proceed to construct such large arrays. Reasonable gains, however, can be obtained with arrays of axial mode helix antennas. The helix is attractive at 1.5 GHz mainly because of its relatively small size. The design shown in **figs. 17A, B**, and **C** indicates that the length of a helical beam at this frequency would be about 19 inches with a 2.5-inch diameter and a minimum reflector size of only 0.8 wavelength. This would make an inconspicuous installation. Although the helical antenna is not known for its gain, it has been used extensively by professional radio astronomers. A nine-turn helix antenna can provide about 14.8 dB, and a twelve-turn *helix*, about 16 *dB of gain*.

Helix antennas have also been used in more moderate arrays with gains in excess of 25 dB. Depending on which way they are wound in regard to each other, several polarization schemes can be accomplished. For example, using a pair of helices with the same sense (both clockwise or vice-versa) can provide circular polarization. Using opposed windings allows for horizontal polarization. A four-antenna array with clockwise and counterclockwise components can be interconnected so that several choices of polarizations could be obtained, as shown in **fig. 17D**. In addition, beamwidths of 10 to 15 degrees have been achieved.

The main characteristic of the helix antenna is its relatively wide bandwidth (− 20 percent and + 30 percent of the center frequency), which makes it suitable for the water hole band. Unless terminated with a matching strip or at a special point on the back plane, a helix exhibits a high impedance output of about 140 ohms. Inasmuch as this could be a disadvantage in a single-antenna design, parallel arrays using high-impedance coaxial cables can produce composite outputs of 75 ohms without the use of RF combiners.

On the other hand, a helix antenna can exhibit a much higher noise figure than the parabolic dish. Because the noise temperature of an antenna is determined by the noise power available in its lobes (this includes its minor lobes), if the antenna is "looking" at the ground — which has a typical noise temperature of 290 degrees K (17 degrees C) — it will have a noise figure of approximately 3 dB, which would be much higher than that of a preamplifier. In this respect the parabolic antenna would be better (lower side lobes). Careful consideration for the location of helical array is recommended; the choice of polarization, explained earlier, can also greatly improve the system. Fourier transforms performed with simple microcomputers will also help in separating the desired components from the noise in these lobes.

conclusion

Although it may be difficult for Radio Amateurs to accept the seemingly impractical nature of radio astronomy projects, the experience gained in developing one's own system could provide a complete education in contemporary radio communication.

Detailed information on the construction of radio astronomy and SETI projects, including low-noise amplifiers, and fast Fourier transform (FFT) programs for simple microcomputers can be obtained from The Society of Amateur Radio Astronomers (SARA) which publishes several books on the subject along with a monthly newsletter. At present, The Society has 168 members worldwide, many of whom are hams as well as scientists and engineers working in related fields. For more information, write to Robert M. Sickels, Secretary, SARA, 7605 Deland Avenue, Fort Pierce, Florida 33451.

There are many arguments about the existence of extraterrestrial intelligence. Some scientists believe that intelligent life exists elsewhere in the universe, while others mathematically analyze the probabilities and conclude that we could very well be the only advanced civilization in our galaxy. While this is a discouraging thought, we cannot rule out the possibility that there may be a few others out there — perhaps many others. Although we have no evidence yet to support the claim that ETI may exist, many scientists have been taking the task of SETI very seriously, and an increased number of receiving stations built by Radio Amateurs would only improve the chance of receiving that first intelligent signal from beyond our own solar system. We know that the laws of physics are the same throughout the universe; an advanced civilization, therefore, regardless of what it used to produce RF energy, would radiate the same kind of RF energy we know here on Earth. Our modern RF technology is now producing receivers with noise figures that approach the limitations of intergalactic noise. The gap has finally been closed; *we can now begin the final search.*

acknowledgements

I wish to thank Garred Giles (ex KP4EPM) of the Arecibo Observatory for providing key photographs taken especially for this article. My appreciation is also extended to Dr. Harry Webb of the College of St. Thomas in St. Paul, Minnesota, for sharing his collection of photographs taken in his many trips to the Arecibo Observatory. I would also like to thank Dr. Robert S. Dixon, W8ERD, Assistant Director of the Ohio State Radio Observatory;

Professor Paul Horowitz, W1HFA, of Harvard University's Physics Department; and Bob Zimmerman, NP4B, of the Arecibo Observatory, for their contributions to this work.

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ham radio

the
weekender

harmonic mixer for VHF signal generation

Lots of useful things that work well can be built without great effort and expense. Unfortunately, until now, a VHF signal generator hasn't been one of them. The cost of parts, the shielding required, and the difficulty of producing a good variable attenuator make designing an inexpensive, easy-to-build VHF signal generator a challenging task indeed.

initial means of generating VHF signals

Even though I was able to use a surplus TS-150 generator to meet my needs from 10 through 400 MHz, I was never fully satisfied with it. Recently when I acquired a pair of Clemens SG-83C solid-state high-frequency signal generators (they're cute and small, and they work well; they'll even work off a 9-volt battery!) at a flea market, my old TS-510 began to look more and more out of place. So the search was on for an up-to-date replacement.

There don't seem to be many bargains among generators that will make it to 400 or 500 MHz. The older units have good precision attenuators with high available output levels up to 480 MHz. But most models are marked by poor stability, excessive signal leakage, and large current consumption. They're contained in large, bulky enclosures, seem to be priced higher than they might be, and depend on exotic, expensive tubes and components. But if you own or have access to a generator in the HF range and want some VHF or UHF coverage without buying yet another generator, you can build this device in just a few hours — for no more than $25.00.

heart of device is diode ring mixer

In recent years the use of diode ring mixers has become increasingly popular in state-of-the-art receiver design. Despite their many advantages, they do present some problems. One is harmonic mixing. Diode ring mixers are LO-driven at high signal levels — typically +7 dBm minimum. These mixers are broadband devices, using wideband transformers and fast hot carrier diodes in their construction; the inexpensive Mini-Circuits SBL-1 DBM used in this article is rated from 500 kHz to 500 MHz at all three ports (see fig. 1). When LO power is applied, the diodes are driven into heavy conduction by the RF, producing LO harmonics well into the VHF range and resulting in undesired mixer products from input signals mixing with the harmonics appearing at the IF output port. This required the use of wideband impedance terminations for the DBM ports, as well as HF filters with known stopband characteristics well into the VHF region if the receiver isn't to sound like an aviary at feeding time!

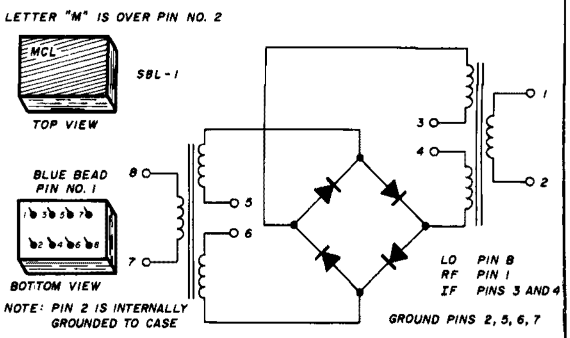

fig. 1. Harmonic generator uses Mini-Circuits SBL-1 mixer.

In this simple adapter circuit we make good use of this phenomenon. Referring to the schematic (see fig. 2), note that we are driving the LO port with the output of a 50-MHz overtone oscillator. The oscillator was designed to produce a minimum +10 dBm output with a 9-volt battery supply. DBMs expect to see broadband resistive port terminations if insertion losses, bandwidth, and other parameters are to remain predictable. The RF port of the DBM is connected directly to the HF signal generator output port; most good generators will present a good 50-ohm resistive termination at their output. This is especially true of the Clemens SG-83 series because of the switched pi attenuators employed to set the signal output level. A 4-dB fixed pi attenuator is used at the DBM IF port output. Further attenuation is desirable here; this will be discussed later in the article.

When the device is used at the output of a signal generator, the 50 MHz LO produces internal harmonics in the DBM at 100 MHz, 150 MHz, 200 MHz, etc. throughout the spectrum. When we inject a signal into our adapter from the HF generator, say at 10 MHz,

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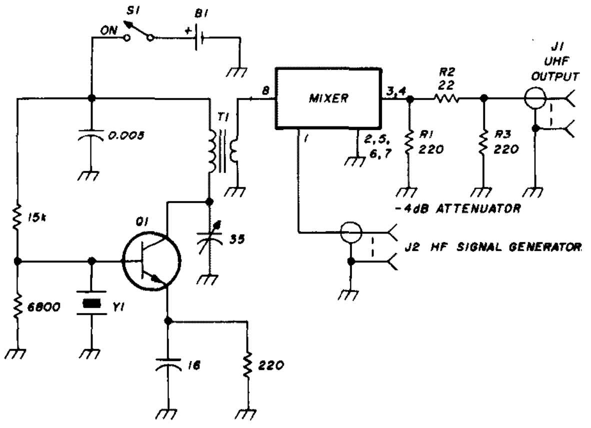

item	description
B1	9 volts, "transistor" battery
J1,J2	BNC; chassis fittings
Q1	2N918 transistor
R1,R3	220 ohm, 1/4 watt resistor
R2	22 ohm, 1/4 watt resistor
S1	single-pole, single-throw miniature switch
T1	primary 12 turns No. 26 wire, secondary 2 turns No. 26 wire
Y1*	3rd overtone 50.0 MHz crystal
mixer	Mini-Circuits Labs (or equivalent) model SBL-1,** 0.5-500 MHz, +7 dBm local oscillator

*50 MHz crystal available from International Crystal Manufacturing Co., 10 North Lee, P.O. Box 26330, Oklahoma City, Oklahoma 73126. Request "general purpose" grade crystal.

**SBL-1 mixer is available from Advanced Receiver Research, Box 1242, Burlington, Connecticut 06013; send SASE for price quote.

fig. 2. Harmonic mixer provides output up to 500 MHz.

outputs at the sum and difference frequencies will result from the 10 MHz signal combining with the 50 MHz LO signal (50 + 10 = 60 MHz or 50 − 10 = 40 MHz) as well as outputs of the sum and difference frequencies of the LO harmonics. From this it is evident that the HF signal generator need only extend to 25 MHz to produce contiguous coverage over the entire range (up to at least 500 MHz) by using either the sum or difference products of the appropriate harmonic. A generator that will go to 50 MHz will allow the output frequency to be determined without involving mental subtraction.

there *are* limitations, though

There is a price to pay for this convenience. First, the conversion losses will vary across the VHF ranges because the LO harmonic amplitudes produced in the DBM are not identical. This is determined by directly comparing the output of the adapter versus the output of a good VHF generator of known calibration into a receiver with signal strength metering for each of the harmonic bands. The receivers should be terminated with an external 6 dB or greater attenuation pad; most VHF receivers present only a reasonable 50-ohm load over a very narrow range of frequencies — (if any).

use is simple

Once the conversion losses are known for the adapter for each of the harmonic ranges, the rest is easy. If you work in the dBm signal levels you need only include the dB loss of the adapter to the dBm reading of the signal generator output. If you prefer to work in microvolts, an easy solution is to employ a separate switched attenuator capable of one dB steps between the signal generator and the adapter. By keeping the adapter losses at 20 dB increments the microvolt scale of the HF generator may be read directly; for each 20 dB of insertion loss the microvolt reading is moved one decimal point to the left.

Spurious responses may be encountered at some desired frequencies when using the adapter. These birdies are easily identified, however, because they tune "backwards" or at rapid rates across the receiver bandpass. Note that when using this adapter, HF signal generator leakage is of minimal importance because it is not the final output frequency. Extremely low MDS measurements can be made, and the conversion loss inherent with this adapter can be advantageous here as some generators (such as my Clemens) are limited to minimum output levels around −120 dBm unless external attenuation is employed.

construction is easy

I built my adapter in a small minibox and used point-to-point wiring techniques on a small square of circuit board, following good VHF techniques such as keeping leads short and using good shielding. I also recommend using the internal battery supply as shown; bringing power leads out of the adapter enclosure is an invitation to stray radiation. The oscillator draws about 10 milliamperes and battery life should be quite long.

The maximum output levels from the adapter are severely limited by the conversion losses, which can be very high on some harmonics, and by the compression that takes place in the mixer if it's overdriven. I've been using my adapter for a few weeks now and I don't know how I ever did without it. The freedom from drift and generator leakage — and ease of setting frequency — are *pure joy*. My old unit is gone.

ham radio

designing low voltage power supplies

Husky power supply delivers 15 amps at 13.8 volts

Power supplies for tube equipment used to be fairly simple devices, easily assembled from parts found in any handy junk box and quite forgiving of design error. But low-voltage regulated supplies for solid-state gear are something else again.

Poorly designed low-voltage supplies can create instability of transmitted signals, distortion of received signals, hum, noise, spurious emissions and transients. A well-designed supply, on the other hand, contributes greatly to clean, trouble-free station performance.

specifications

At the heart of any power supply is the transformer, which takes energy from a commercial power line and converts it to some specified lower or higher voltage at a specified current. The single most expensive unit in the whole assembly, the transformer is usually called upon to absorb more abuse than any other component. Once you've specified the transformer, you've almost specified the rest of the supply.

Because the transformer is the most expensive component, it's also the one in which manufacturers may tend to cut corners. Sometimes they'll put a 2-ampere transformer in a box with a few other components and advertise the result as a "5-ampere power supply."

choosing the transformer

The three principal specifications for the transformer are primary voltage, secondary voltage, and maximum secondary current. To specify the secondary requirements, you must first decide how much filtered DC current and voltage you want. For the sake of illustration, let's walk through the process of designing a supply that will deliver about 15 amperes at 13.8 volts continuously and about 20 amperes intermittently, since that sort of supply would handle most ham station requirements.

Fortunately, most manufacturers rate their transformers for continuous service. Transformers are bulky objects with a lot of thermal inertia; they can absorb a lot of heat without getting hot too quickly. Consequently, a transformer rated at 10 amperes continuous duty can probably deliver about 14 amperes or more on a 50-percent duty cycle and perhaps even more current on a shorter cycle without serious risk.

Perhaps the most severe duty most ham stations impose on a power supply (aside from beacon or repeater duty) is a long-running activity such as the ARRL Sweepstakes or the *CQ* WW contest. In contests, operators listen at least as much as they transmit, automatically setting the duty cycle at 50 percent or less. An overloaded transformer can rest and cool off during the time it is not required to deliver power to the transmitter.

If you're operating CW, the key will be down and the transmitter putting out only about half the time you're in the transmit mode, so your net duty cycle is down to 25 percent. And if you're working SSB, the duty cycle is even lower.

So if an HF rig demands 20 amperes key down,

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what's needed is a power supply that will handle about 25 percent of that or 5 amperes continuous, right?

Wrong!

Heat generated by internal losses is the biggest enemy of transformers. These losses rise steeply as the current demand exceeds capabilities. The losses have several causes:

- The ohmic resistance of the windings is a major source of heat generation, especially at high currents, because heat (power) equals I^2R. When you double the current in a winding, you quadruple the heat. As copper gets hotter, its resistance increases still further.
- Eddy currents, induced in the iron core of the transformer, generate heat. The larger the current in the windings, the larger the eddy currents. These losses can be minimized by careful design of the transformer by the manufacturer.
- Hysteresis losses also generate heat. They result from the use of some transformer energy to jostle iron atoms in the core, magnetically exciting them. Simply put, energy is lost in rearranging the magnetic patterns in the core.
- Dielectric losses occur in a transformer because it is inherently a capacitor as well as a transformer; it is several masses of metal in close proximity, separated by insulation. The insulating dielectric is usually quite lossy; heat is generated by jostling its molecules around in the rapidly-changing electrostatic fields inside the transformer.

This means that you *can* run a transformer harder than its continuous-duty rating, but only within certain limits. Even if you could keep the transformer cool, hysteresis, eddy current and ohmic losses would severely degrade performance.

Generally, however, it is reasonable to expect about 40 percent more — at most — than the continuous rating. So, if a rig wants 20 amperes key down, we can get away with using a transformer rated at 12 amperes continuous duty. But a 15-ampere rating would be safer and cooler and would pose a smaller risk of failure.

overall design

Since abrupt changes in supply voltages can cause trouble in electronic gear, a power supply must be stiffly regulated, varying output voltage 0.1 percent or less when the load goes from open circuit (no current) to the design limit of the supply (20 amperes). Such regulation is easily obtained with inexpensive parts. In addition, we must have DC with the AC ripple component reduced to an insignificant level.

Because both cost and performance are concerns, this design will be fairly conservative, without much consideration of size and weight.

For simplicity, we'll use a conventional series-pass transistor circuit controlled by an electronic regulator. That means that the DC output from the filter will have to be somewhat higher than the desired regulated voltage. Most electronic regulators require 2 or 3 volts difference between input and regulated output. Since we want 13.8 volts regulated, we'll need at least 16.3 volts out of the filter at full load for reasons that will soon become apparent.

One rule of thumb says that under load, a power supply will deliver a DC output about equal to the applied RMS voltage for a full-wave bridge, or about half the total secondary voltage in a center-tap circuit. Actually, 90 percent of the RMS voltage is a bit more accurate if the transformer is taxed to its full load limit. Thus, an 18-volt secondary will deliver about 16 volts DC ($0.9 \times 18\ V_{RMS}$) under full rated load. But with no load, the DC output voltage will approximate the AC peak voltage, which is $1.414 \times V_{RMS} = 25.5$ volts. Using the same formula, you can see that a 30-volt secondary will deliver about 42 volts at no load — too high for our regulator chip, whose maximum input rating is 35 volts. The maximum RMS output we can tolerate will be $35\ \text{volts} \div 1.414 = 25$ volts.

So our transformer must have a secondary that gives us between 19 and 25 volts AC RMS at about 15 amperes continuous. If the transformer is center-tapped, for use with only two diodes, we need that much voltage on either side of center. If we're using a bridge, we need only a single winding delivering the desired voltage.

Let's assume that our junk box contains a 12.6-volt filament transformer rated at 15 amperes, and that we can scrounge a 6.3-volt filament transformer rated at 20 amperes. Connecting primaries in parallel and secondaries in series so that their voltages add, we get 18.9 volts RMS AC. The amount of current we can draw at that voltage is limited by the lower rating — 15 amperes. But because the output voltages from the two secondaries will not be precisely in phase (because of differences in manufacturing), we must derate their current capacity by about 10 percent. So we can demand only 90 percent of 15 amperes, or 13.5 amperes from the power supply on a continuous basis, or about 19 amperes intermittently. Given the uncertainties involved in estimating duty cycles and the amount of uprating and derating, that should be close enough.

So we can expect to get about 17 volts at full load and about 27 volts at no load from the rectifier-filter output.

selecting rectifiers

Because we chose a transformer without a center-tap, we'll need to rectify the AC with a full-wave bridge. Radio Shack and other suppliers offer an in-

expensive bridge rated at 25 amperes at 50 volts peak inverse voltage (PIV).

In a bridge circuit, the full transformer secondary voltage appears at the nonconducting anodes. Since our transformer secondary voltage is 26.9 volts peak, the 50 PIV rating (also sometimes called the peak *reverse* voltage, or PRV) is quite satisfactory, but not extreme. Discrete diodes could also be used in a bridge circuit, of course, with each rated to handle at least half the total load current with a PIV above the peak output voltage of the full secondary winding of the transformer. For this supply, diodes rated at 50 PIV and 10 or 15 amperes each would be satisfactory.

choosing the regulator

In any power supply the filter immediately follows the rectifiers. But in a regulated power supply, the choice of regulator circuit is intimately connected with design of the filter for reasons that will soon become clear.

The regulator samples the output voltage of the power supply, compares it to a fixed reference voltage, detects any difference, and applies a correction voltage somewhere in the circuit to counteract any change in output voltage.

The earliest and simplest regulators used a zener diode for a reference and simply applied that voltage to the base of a pass transistor, clamping the output at approximately the zener voltage (**fig. 1**). This circuit was quite a big improvement over the zener diode alone for two reasons: first, as a series, rather than shunt regulator, it could handle a lot more current; second, it offered a rudimentary sort of "electronic filtering." If a capacitor is connected between the base and ground, its filtering effect (capacitance) will be multiplied by the β of the transistor. Thus, if the transistor is a 2N3055 with a typical β of about 70, it would make a 100 μF capacitor do the filtering job of a 7000 μF capacitor.

The disadvantage of this circuit is that it provides no feedback loop. When power is taken from the emitter, a voltage drop occurs across the silicon emitter-collector junction and across any DC resistance in wires and connectors. Such a "regulated" supply's output may easily drop several volts under a heavy load. That's not good regulation!

Another circuit frequently seen even now in magazine articles is basically the same as above with an IC regulator chip substituted for the zener (**fig. 2**). It has exactly the same disadvantage as the zener circuit — no feedback loop. The IC chip will hold the base voltage very tight but it has no control of the voltage at the load.

Another circuit, using a zener diode and discrete components, is a considerable improvement over the others (**fig. 3**). In this circuit, the zener serves as a reference, and the base of a regulator transistor samples the output from the supply. Any change in the ratio between the zener reference voltage and the base voltage is converted into an error signal at the emitter and is applied to the pass transistor.

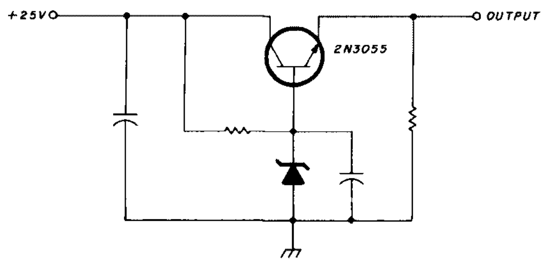


fig. 1. Basic voltage regulator using a zener diode and pass transistor. Disadvantage of this circuit is that it provides no feedback loop.

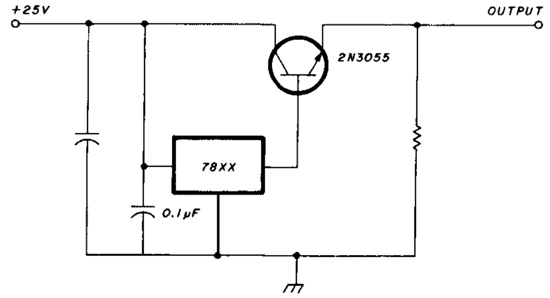


fig. 2. Regulator circuit using an IC chip in place of the zener. Circuit also has no feedback loop; IC chip regulates the base voltage but has no control of overload voltage.

A further embellishment of this idea (**fig. 4**) is to apply the error voltage to a driver transistor that also serves as an error amplifier, Q2, resulting in high sensitivity and good regulation. The regulator transistor, Q1, can be quite small (a 2N2222, for example) but the driver must be able to handle the base drive current required by the pass transistor, Q3. With suitable transistors, this circuit should be able to maintain load regulation of better than 1 percent.

three-terminal regulators

Properly used, however, the integrated circuit regulator chips can put us into a whole new ball park of regulation — 0.1 percent or even 0.01 percent with off-the-shelf parts.

The circuit in **fig. 5** shows a 78XX regulator with

a PNP "wrap-around" pass transistor.[1,2] The regulator chip and the pass transistor share the load current in a ratio set by resistors R1 and R2. Any change in the load voltage is sensed internally by the chip and applied to its input where it is also applied to the base of the pass transistor. The diode simply increases the voltage differential between input and output by the amount of the silicon junction voltage drop.

With the proper pass transistors, this circuit can deliver almost any practical amount of current. To use NPN transistors requires a PNP driver stage between the regulator chip and the bases of the pass units to provide the correct bias voltage polarity. This sort of circuit is shown in **fig. 6**.

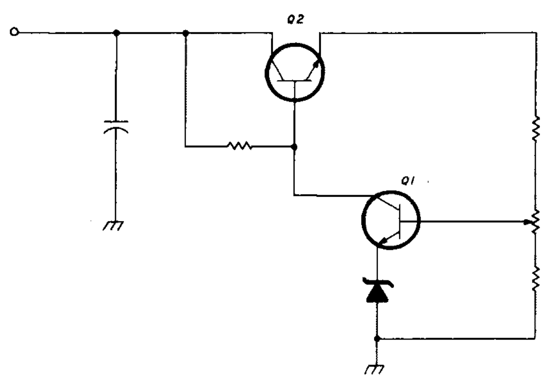


fig. 3. Regulator circuit in which the zener is a reference and base of regulator transistor, Q1, samples supply output.

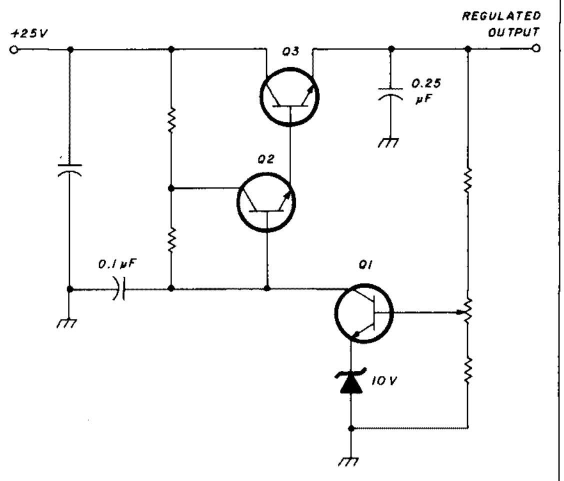


fig. 4. Improved regulator uses a 2N2222 regulator transistor. TIP-31 driver-error amplifier, and 2N3055 pass transistor. Regulation better than 1 percent can be obtained.

Both the circuits of **figs. 5** and **6** produce voltage regulation on the order of 0.1 percent. The more sophisticated LM-723 regulator chip, with a higher parts count required in its circuitry, is capable of regulating a whole order of magnitude better than the 78XX regulators; about 0.01 percent is typical. The circuit shown in **fig. 7** is representative, but there are some trade-offs.

The 723 can deliver only about 150 mA at the regulated output voltage and so requires a driver stage for most pass transistors. It also requires a rather long parts list and, unless carefully shielded and bypassed, is often subject to RF interference that causes it to operate erratically. One redeeming feature is that its voltage-sensing connections are isolated and can be connected directly to the load by a long run of small wire. This permits the regulator to compensate for voltage drops in the wiring to the load.

Like the 78XX chips, the 723 offers automatic shutdown if it gets too warm or if excessive current is drawn from the supply. A short circuit across the output terminals of even a 50-ampere supply produces only a minor spark and a turn-off of the voltage. The voltage recovers to normal when the short circuit is removed, with no harmful effect. Over-voltage and over-current protection circuits are discussed later in this article.

After considering all the possibilities, we have decided to use the circuit of **fig. 6**, permitting us to use NPN transistors, inexpensive, readily-available regulator chips, and a minimum of external parts.

the 7812 regulator

Many good regulator IC and discrete-component circuits are available to Amateur builders and many of them offer excellent regulation. The LM-723 offers 0.01 percent regulation of both line and load, but at the expense of fragility and added circuit complexity.

Line regulation is the percentage of output voltage change with a change in power line voltage. Load regulation is the percentage of output voltage change with a change in the load current.

For virtually all but laboratory purposes, 0.1 percent regulation is more than adequate and is readily attainable with simple circuits and inexpensive components.

Since we're building what amounts to a fixed-voltage supply, we can eliminate most of the IC regulators that provide adjustable output such as the LM-317, the LM-350, and the LM-723.

Members of the 78XX series, on the other hand, are adjustable within limits, can deliver more than an ampere of regulated output, require very few external components, and provide regulation better than 0.1 percent. (The 7812 regulator is used in our finished power supply.) In addition, they have internal short-

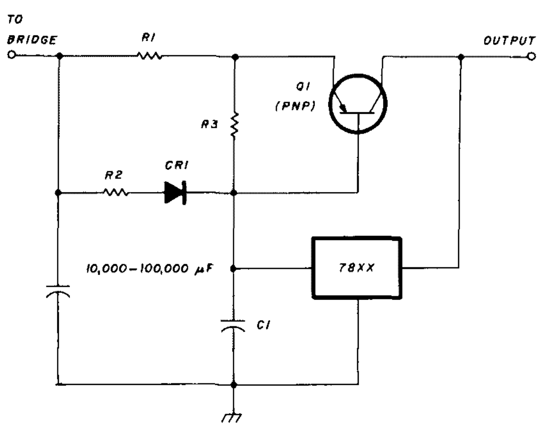


fig. 5. A 78XX regulator with a PNP "wraparound" pass transistor. Regulator IC and pass transistor share load current in the ratio set by R1 and R2.

circuit protection, internal over-heating protection and can lend these properties to the pass transistors as well. These chips have a ripple rejection of better than 50 dB and require an input only about 2.5 volts above their regulated output. The regulated output voltage is adjustable within limits by inserting a resistor in the ground lead. That's a lot of regulator for a couple of bucks!

designing the filter[1-3]

The requirement of a 2.5-volt differential between input and regulated output sets the dropout voltage. That's the input voltage below which the regulator chip loses control. At the dropout voltage, any further reduction in the input produces a similar reduction in the output.

The dropout figure is the reason we deferred design of the filter until the regulator was chosen. What does that have to do with filter design? Plenty. Any AC ripple voltage on the regulator input makes the instantaneous voltage rise and fall at a 120-Hz rate. The larger the ripple voltage, the greater the peak swings of the input voltage. If the input dips below the dropout value on those negative peaks, the regulated output also drops, introducing ripple into the output.

Obviously, we need filter design that will keep the negative ripple peaks above the dropout voltage.

Since we want an output voltage of 13.8 volts, we'll select a 7812 regulator chip and raise its voltage with a resistor in the ground lead. And because data sheets for the 7812 put the input dropout voltage at about 2.5 volts, we must supply an input voltage of at least 16.3 volts; that is, *the input voltage must never fall below 16.3 volts* or ripple will appear in the output.

We have already determined that the output voltage of the rectifiers and filter will be 17.0 volts under full load. How much ripple can we tolerate without going into the dropout zone? The difference between 17.0 volts and 16.3 volts is 0.7 volt, which is the value of the negative *peaks* we can tolerate. That would give us 2 × 0.7 volt or 1.4 volts peak-to-peak, as the maximum tolerable amount.

How much capacitance do we need between the bridge output and ground to limit ripple to this amount? All recent editions of the ARRL *Radio Amateur's Handbook* explain the problem in detail. In short, for a ripple frequency of 120 Hz, the required capacitance in μF would be:

$$C_{\mu F} = \frac{I \times 8.3 \times 1000}{E}$$

where I is the load current and E is the peak-to-peak ripple voltage. In our supply, I = 20 amperes and E = 5.2 volts. The required capacitance would be 118,571 μF.

As a matter of economics, a commercial manufacturer might choose a 120,000 μF capacitor or choose to use a higher input voltage to reduce the required capacitance. The purpose of the design computation, of course, was to determine the *minimum* permissible amount of filtering. But a savvy Amateur would use more than the minimum — probably at least 150,000 μF. Such values are easily obtained by paralleling computer surplus units purchased at hamfests or from mail-order suppliers or obtained from a friend's junk box.

Since the maximum amount of allowable ripple

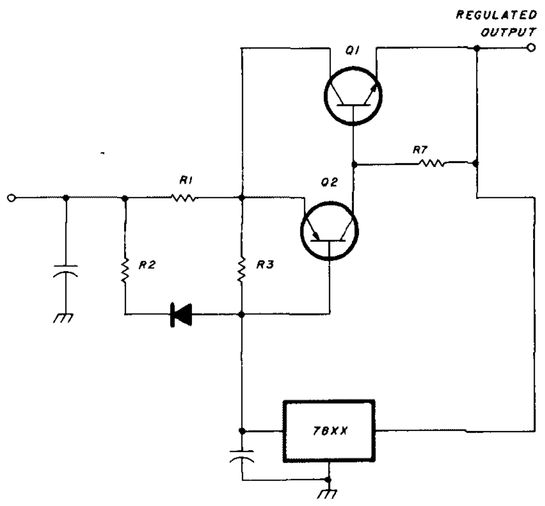


fig. 6. Another version of the 78XX regulator using an NPN pass transistor. A PNP driver stage provides correct bias voltage polarity. Voltage regulation of this circuit and that of *fig. 5* is of the order of 0.1 percent.

determines the amount of filter capacitance required, the problem could also be attacked by increasing the input voltage to the regulator, thus increasing the allowable ripple level. This can be done simply by using a transformer with a higher output voltage, with the trade-off that the pass transistors will have to dispose of more excess voltage, thus dissipating more heat.

A more sophisticated way to get the best of both methods is used in some Astron commercial power supplies, among others. An example appears in **fig. 7**. Here additional taps on the transformer secondary provide higher voltage at low current to operate the regulator chip while the lower-voltage, high-current winding operates the pass transistors and supplies the main output.

In the circuit shown, the regulator chip (an LM-723) takes its input from the higher-voltage supply that is separately rectified and separately filtered. Because the load impedance on this supply is quite high (small load current compared to the input voltage) it is quite easy to filter well, and the higher voltage input allows even greater leeway with the ripple content.

The highly-filtered input to the regulator chip could also be provided by a separate transformer, avoiding the need for a specially-wound transformer such as those found in some commercial units.

Separate regulator supplies cannot be used with the 78XX wraparound regulator circuit we have chosen for our example supply, however.

choosing a driver stage

In our chosen circuit, we need a PNP driver between the regulator chip and the pass transistors to invert the polarity of the error-voltage changes. In other circuits we might also need a driver to supply sufficient base drive to the pass elements, as in **fig. 7**, for example. But how much drive current do we need?

The answer lies in our choice of pass transistors. We will use two 2N3055s to deliver 20 amperes. The β of these transistors typically is about 70. Twenty amperes divided by 70 is 0.286 ampere, which is the amount of drive current we need. But to be on the safe side, we'll choose a transistor that will deliver at least twice that amount. It happens that an ECG 129 will handle 1 ampere of collector current and has a β of about 100. It is an inexpensive, readily-available PNP transistor and so would make a good choice.

In heavier supplies, or when it is necessary to drive less sensitive pass transistors, thus requiring more base drive than 1 ampere, it will be necessary to use a heftier driver transistor. The Darlington PNP IC designated TIP-125 will deliver up to 5 amperes of drive current and, because it has a β of at least 1000, the drive it requires from the regulator is very small indeed. Thus the TIP-125 could be used with such regulators as the LM-723 and the 78LXX series, which deliver only about 100 mA of regulated output.

choosing pass transistors

Pass transistors are the "valves" that control the flow of current from the rectifiers to the load, and they act much as though they were variable resistors in series with the load. They must absorb all the power supply energy not taken by the load. In our case, they must absorb a voltage drop of 3.2 at 20 amperes under full load = 64 watts.

For our pass transistors, we'll select the readily-available, rugged, and inexpensive 2N3055 NPN silicons rated at 115 watts each and 15 amperes maximum collector current. They have a forward current transfer ratio (h_{FE} or β) of about 70. They cost less than $2 each at discount houses and hamfests, and often sell for a fraction of that.

One 2N3055 would handle the full heat load from our power supply — only 64 watts — but one transistor can't handle the 20 ampere maximum current we want to draw.

We can overcome this by connecting two pass transistors in parallel, providing 30 amperes capability. But they can't simply be connected in parallel because all semiconductor devices are slightly different: one transistor, with the higher β, would try to take all the current, leaving little on none for its mate. This is easily corrected with low-value resistors in each collector load, allowing the current to equalize. Power dissipated in each 0.1-ohm resistor equals I^2R where I = half the total load current (the current through each transistor). Since in this case $I = 10$, $I^2 = 100$ and $P = 10$ watts.

However, it probably is not necessary to use 10-watt resistors because $I = 10$ only when the key is down or voice peaks are maximum. The *average*, even in CW, will probably be about 5 amperes. At that level $I^2 = 25$ and $P = 2.5$ watts. Most likely a 5-watt resistor will work fine. Resistors, like transformers, have a relatively large thermal inertia because of their mass — a wholly different situation than prevails in semiconductors, which will fail immediately when their maximum ratings are exceeded.

Exact resistance values are not critical in the emitter-equalizer applications so long as they equal each other and offer a very low voltage-drop at maximum current. Two 5-watt resistors in parallel could be used in place of a single 10-watt resistor.

heat sinking[3-11]

Transistor junctions, even in big power transistors, are very small devices: they have little mass, hence low thermal inertia. In other words, they get hot almost instantly, even in normal use. Power transistors are packaged in cases designed to dissipate this heat; they partly make up, by their own bulk, what the silicon junction lacks in mass. That's why the collectors on most power transistors are connected to the metal case of a TO-3-type package and to the tab of a

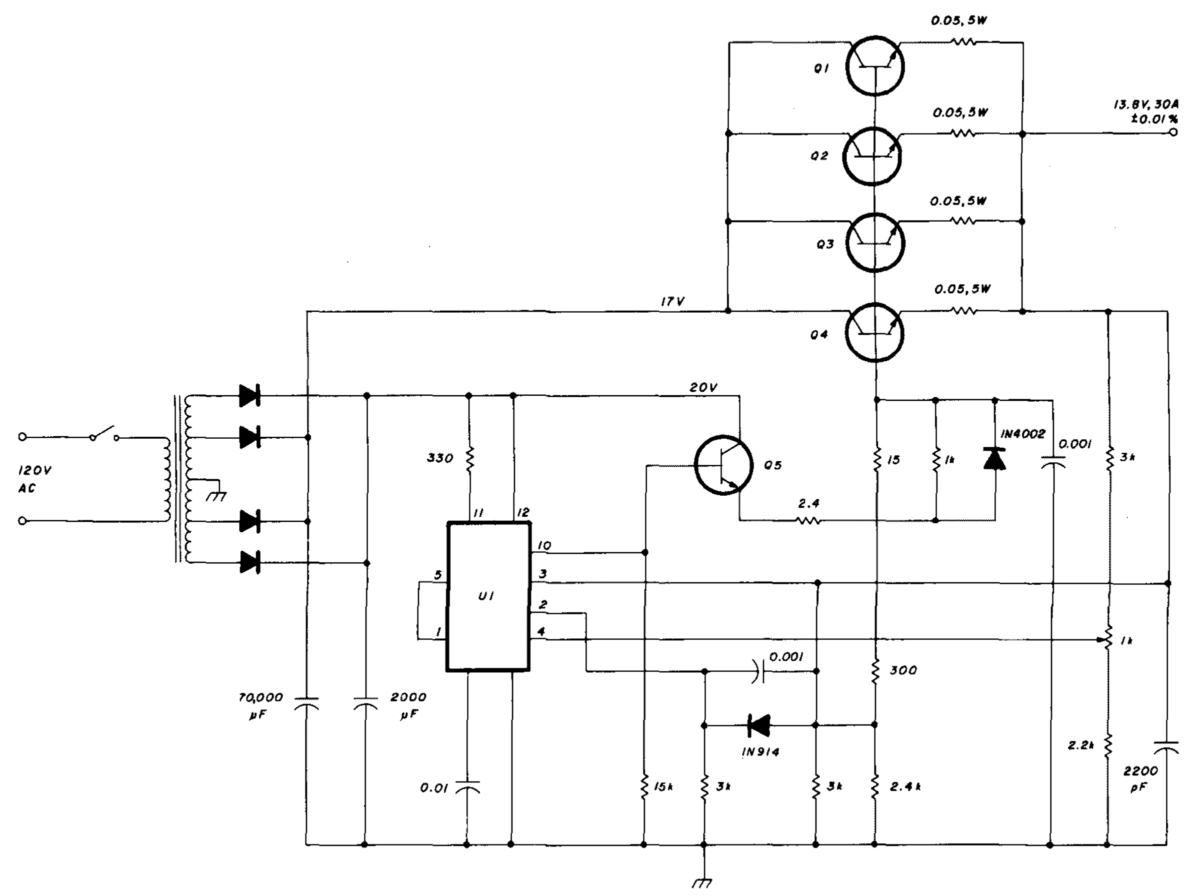


fig. 7. Simplified diagram for Astron RS-35M power supply.

TO-220-type package. The collector generates most of the internal heat because it must carry all the emitter current plus all the base current.

External heat sinks are large masses of metal, such as aluminum or copper, that are highly conductive of heat and have a large surface area to help transfer that heat to the surrounding air. A lot of complex formulas exist for calculating the necessary amount of heat sinking.[3-11] They take into account the mass of the heat sink, the temperature of the surrounding air, and the degree of thermal coupling between the transistor case and the heat sink.

Use these formulas if you like, but a Sylvania shop note offers a better practical suggestion: "Use the biggest heat sink you can find that will fit into the space available."

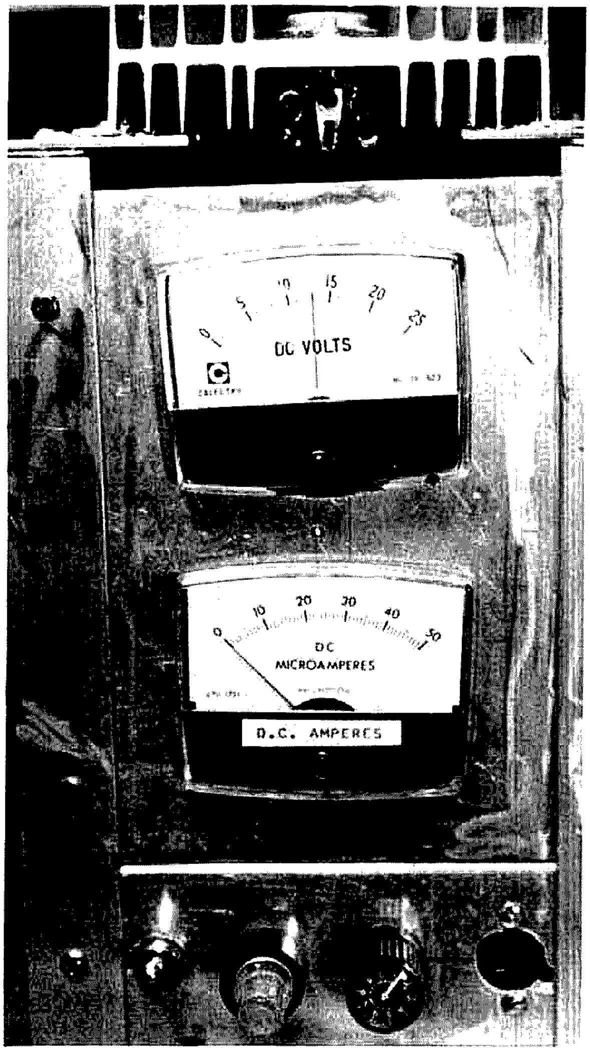
The pass transistors obviously need a large amount of heat sinking because they must dissipate a large amount of heat at maximum current. So do the rectifier diodes or bridge.

The regulator chip and driver transistor will also perform better and last longer if they are used with an adequate heat sink. Power resistors, used in several places in the circuit, don't need heat sinks, but they should be mounted on top of the chassis where air can circulate around them.

stabilizing the regulator

Regulators, whether they are IC chips or discrete components, are high-gain amplifiers and thus are subject to self-oscillation unless precautions are taken to prevent it.

Sometimes such suppression is quite easy and sometimes it requires a great deal of experimentation. But no regulated power supply should be put into service until it has been given a thorough test for oscillation, preferably using an oscilloscope connected across the output terminals of the supply. Oscillations, usually in the 10 kHz to 10 MHz range, will appear as thickening of the trace even if they are not resolvable into individual cycles. Another test option would be to use an RF probe with a VTVM. The meter should



Front panel of 50-ampere power supply. Power ON/OFF switch is at bottom left. Fuse holder is next (with neon lamp to indicate blown fuse). A 10-turn pot is used to adjust voltage. A 2-pin Jones plug (lower right) provides output voltage for small loads.

show no output at all if the power supply is stable, and will give readings of probably several volts if oscillations are present.

The supply may be stable at no load, but break into oscillation at some values of load current while remaining stable at other values. It is necessary to apply several different loads while checking for oscillations to be sure the supply is stable. A good range would be 10 percent, 50 percent, and 110 percent of the rated capacity of the supply.

A standard precaution to observe during construction is to return all ground leads to a single point in order to avoid ground loops. A suitable point would be the negative terminal of the filter capacitor. This would be connected to the chassis by a heavy conductor leading directly to the output terminal and grounded there. This connection must be made with wire heavy enough to handle the entire rated output of the power supply without significant voltage drop. Two or three pieces of flexible copper braid from RG-8 coax would be suitable for a 30 or 40-ampere supply. Cables sold for automobile battery connections are also usually adequate.

All 78XX regulators must be bypassed directly from the input pin to the control pin by a 0.1 μF capacitor connected right at the terminals of the device. Ordinarily, this is enough; however, it sometimes occurs that the inverter/driver transistor or the pass transistor(s) or combinations of both will oscillate. This is especially true if leads are longer than an inch or two between these devices and the IC regulator or between transistors. In most supplies, these leads are much longer than that because of the sheer physical size of the components.

The cures are usually found by cut-and-try. Try bypass capacitors of 0.01 to 1.0 μF. Disc ceramic, metal film, solid tantalum, electrolytic or Mylar are suitable, but capacitance values of electrolytics should be in the 2-10 μF range. Try first bypassing base connections right at the transistors. In a parallel-pass-transistor circuit, try bypassing one transistor but not the others, thus unbalancing any push-pull oscillation.

If bypassing bases doesn't work, try bypassing emitters or collectors. If necessary, insert 100-ohm resistors in series with base connections and bypass one end of the resistors. But don't use resistors in high-current leads, or substitute driver or pass transistors of lower β, for obvious reasons.

The supply may be considered stable if it generates no oscillations at any load within its rated output, even when the load is switched on and off rapidly and repeatedly.

RF sensitivity

Another potential source of instability in a power supply is sensitivity to strong RF fields produced by local transmitters. This is not self-oscillation, but it can exhibit some of the same symptoms — loss or degradation of regulation. With either external RF or internal oscillations, it is not uncommon to see the voltage *rise* when a load is applied.

The cure for this problem is to keep RF out of the power supply. It is most troublesome at the very low-level stages — the regulator chip or transistor. In extreme cases, enough RF may be present to upset a pass transistor or driver.

Enclosing the power supply in a metal cabinet is simple and usually the only RF protection necessary, especially if the cabinet is connected to the station ground. The next simple, obvious step is to bypass AC leads to the chassis and to each other where they

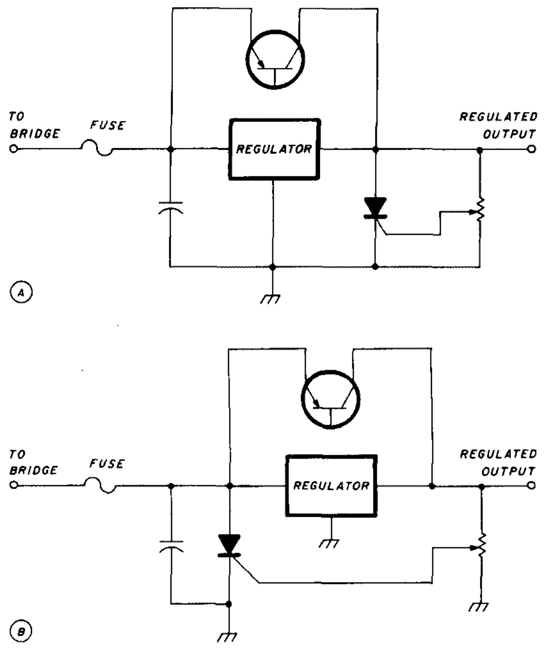


fig. 8. Crowbar circuits using SCRs. In (A) if the regulator fails, the overvoltage will continue to pass excessive current to ground with possible destruction of the SCR, transistors, and transformer. In (B), the SCR is connected across the filter capacitor so that if an overvoltage occurs the short circuit merely blows the fuse.

enter the cabinet. The core of the transformer should also be grounded, along with its internal Faraday shield (if it has one). Suitable AC line bypasses would be 0.02 to 0.05 μF at 600 volts or more.

The output should also be bypassed by capacitors of 0.05 to 0.25 μF because the leads taking power to the load can act as an antenna, feeding RF back into the supply. In aggravated cases such as might occur when operating with a high VSWR, with high RF voltages all over the station, it may be desirable to enclose the regulator chip in an RF-tight metal box within the power supply cabinet, with leads entering and leaving through 0.005 μF feed-through capacitors.

Such extreme measures will seldom be necessary, but it is well to be aware of available curative methods should problems arise.

overvoltage protection[12-19]

If a regulator chip were to fail or if a pass transistor were to short, the result would be that the full output of the bridge would appear at the output of the power supply, with dire consequences for any 12-volt solid-state equipment connected to it.

One common means of protection against this potentiality is a "crowbar circuit." Its usual arrangement includes a silicon-controlled-rectifier (SCR) rated to handle at least the full rated output current of the power supply. The trigger of the SCR is connected to the wiper of a potentiometer connected across the regulated output of the supply. The potentiometer is adjusted so that if the output voltage rises above some predetermined value — usually about 15 volts — the SCR is triggered into conduction, becoming, in effect, a short circuit or "crowbar."

In many commercial supplies, the crowbar is applied directly across the regulated output of the supply. The theory is that this will actuate the automatic overcurrent shutdown feature of the regulator chip, thus shutting off the supply.

It's an excellent theory and it works fine in practice as long as the regulator chip works. But if the primary cause of the overvoltage is failure of the chip, there's a good chance that *it* won't shut down, either. The result is that the overvoltage continues to jam excessive current through the SCR to ground, with probable destruction of the pass transistors, possible destruction of the SCR and possible damage to the transformer itself (see **fig. 8A**).

A much safer procedure is to connect the SCR across the filter capacitor, following a fuse or circuit breaker rated slightly above the maximum allowable current from the supply. Then, when an overvoltage triggers the SCR (with the trigger still connected to the pot across the regulated output), the short circuit merely blows the fuse (**fig. 8B**). Note that if the filter capacitor should short, it would also blow the fuse, thus preventing damage to the transformer.

Another approach to limiting overvoltage is simply to put a big zener diode across the regulated output of the supply. The zener is rated at the maximum voltage you want to allow (for example, 15 volts) and should be designed to handle about as much current as the power supply is rated for.

If a heavy-duty lead-acid battery is floated across the power supply output in parallel with the load, it alone will limit overvoltage to reasonably safe values. If the regulator fails or a pass transistor shorts, the excess energy is absorbed by the battery and converted into heat, letting the voltage rise only a little. Prolonged operation in this mode, of course, will eventually damage or destroy the battery, but no major harm is done during short periods, allowing ample time to shut down the supply and the equipment connected to it.

metering

A power supply — especially one designed for heavy duty — is much more convenient if it is provided with meters to read both voltage and current. This can be done with ordinary 0-50 μA meters or even 0-1 mA meters, as follows.

Since 50 μA meters are available inexpensively from Radio Shack and other suppliers, and since they offer

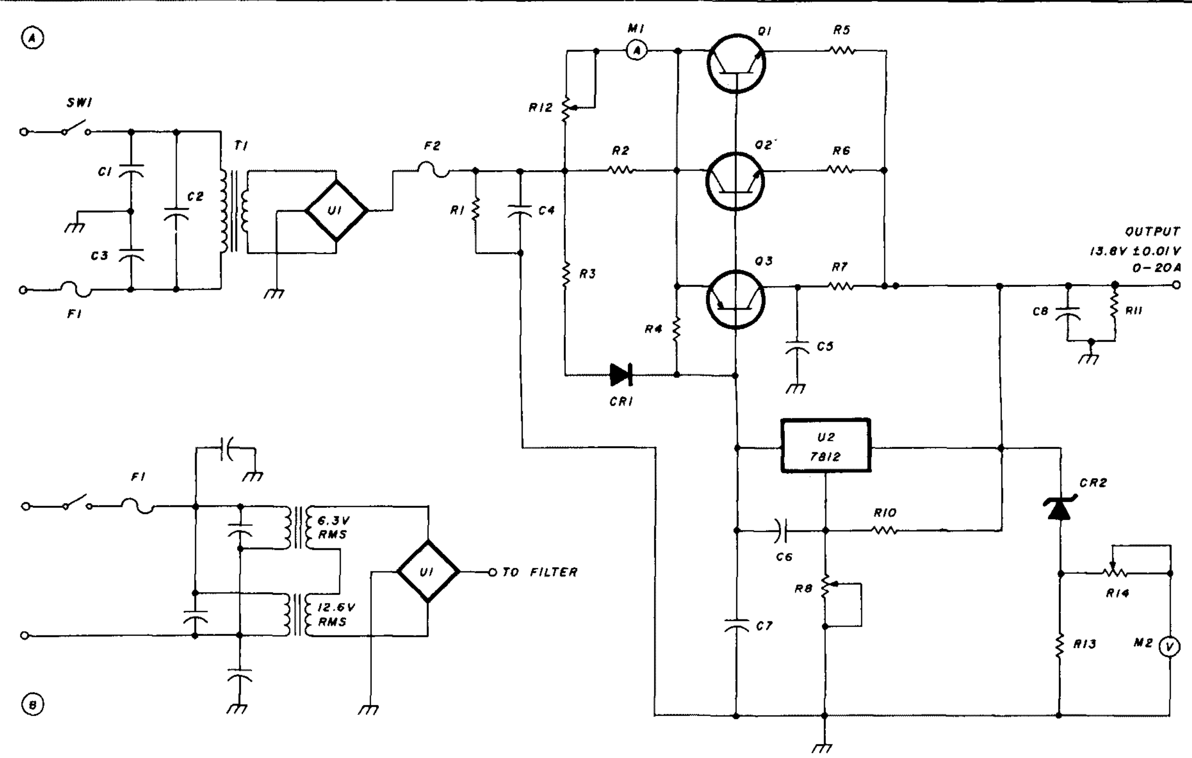


fig. 9. **(A) Schematic of the power supply based on features discussed in the text. Circuit provides 13.8 volts at 20 amperes intermittent, 15 amperes continuous. Regulation is 0.01 volt; (B) Detail of T_1, using transformers in series.**

a convenient scale, we'll use one to measure both voltage and current, as illustrated in fig. 9. Essentially, both circuits use the meter in series with a multiplying resistor to measure the voltage drop across a resistor. This avoids any need to calculate and wind super-accurate, super-low resistance shunts for the ammeter.

To measure voltage, the circuit provides an expanded scale that reads from 10 to 15 volts. CR2 is a 10.0 volt zener selected especially for accuracy. If it operates at any other voltage — even 0.1 volt more or less than 10 volts — the meter will be accurate only at the voltage at which it was calibrated. R13 is chosen to let the zener draw a modest amount of current; a few milliamps is plenty. The μA meter, in series with R14, then reads the voltage drop across R13 which will always be 10 volts less than the output voltage. To calibrate, put an accurate voltmeter across the power supply output and adjust R13 until the meters agree. Ten volts will now be 0 on the meter and 15 volts will read 50.

To read voltage, mentally insert a decimal point and add 10. Thus, 14 volts will read 40 on the scale, and 13.8 volts will read 38. (If you prefer, you can remove the plastic cover from the meter and renumber the dial.)

item	*description*
C1,C2,C3	*0.05 μF 600 volt*
C4	*40,000 μF 35 volt*
C5	*0.1 to 2.0 μF 50 kilohm (see text)*
C6	*0.1 μF 50 volt*
C7	*0.68 μF 50 volt (see text)*
C8	*0.25 μF 600 volt*
CR1	*2.5 amp diode 50 volt minimum*
CR2	*10.0 volt zener, 1.0 watt (see text)*
F1	*5 amp fuse*
F2	*25 amp cartridge fuse*
M1,M2	*0-50 μA DC*
Q1,Q2	*2N3055 or equivalent*
Q3	*ECG 129 or TIP 125 or similar PNP*
R1	*250 ohm 20 watt*
R2	*0.025 ohm 20 watt*
R3	*0.5 ohm 2 watt*
R4	*100 ohm 2 watt*
R5,R6	*0.1 ohm 5 watt*
R7	*75 ohm 1 watt*
R8	*1 kilohm linear*
R9	*omitted*
R10	*1.5 kilohm 1/2 watt*
R11	*150 ohm 5 watt*
R12	*20 kilohm circuit board potentiometer*
R13	*100 ohm 1/2 watt*
R14	*100 kilohm circuit board potentiometer*
SW1	*toggle switch SPST*
T1	*120 volt to 18-25 volt, 15-20 amp*
U1	*25 amp 50 volt rectifier bridge*
U2	*7812 regulator*

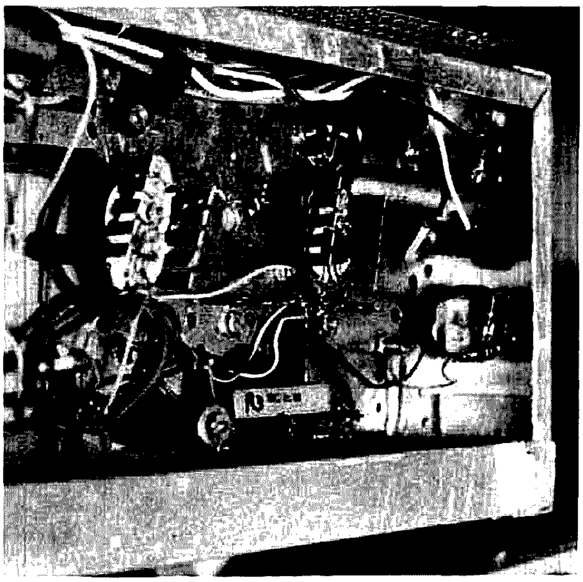
The ammeter, shown as M1 in **fig. 9**, is also rigged as a voltmeter, with R12 as a multiplier. The more current demanded from the supply, the larger the voltage drop across R2. The voltage is directly proportional to the current.

To calibrate, apply a resistor of known value across the power supply output and calculate the current it draws at the output voltage. A 2-ohm load resistor will draw 7 amperes at 14 volts (since I = E/R). A 10-ohm resistor will draw 1.4 amperes at 14 volts.

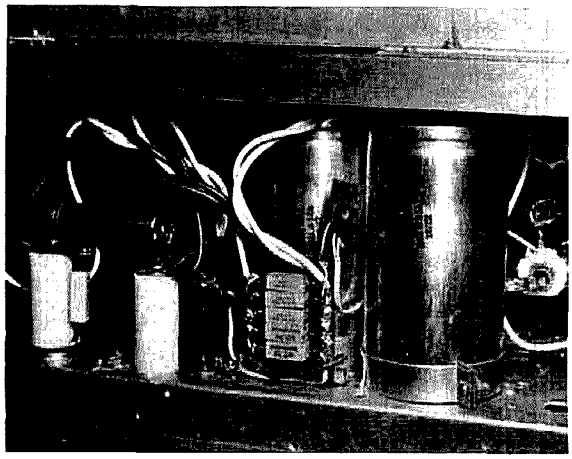
Before you calibrate, however, you must decide what scales you want to use. One scale should have the supply's rated output about mid-scale. Thus, our 20-ampere intermittent duty supply should have one scale that reads 0-40, or more conveniently, 0-50 amperes. Another range could be 0-5 amperes. In either case, the markings on the Radio Shack 0-50 μA meter are appropriate.

To calibrate for 0-50 amperes, use the 2-ohm load resistor and adjust R12 to make the meter read 7.0. The same meter can be used on two ranges by hooking up another potentiometer and switching the positive pole of M1 from R12 to the new potentiometer. The second pot could be set with the 10-ohm load resistor, adjusting it to give a reading of 1.4 (if the output voltage is 14).

The same meter can also be switched, of course, to read both current and voltage.



Bottom view of power supply chassis. Large can-like object at left is the 7 μF non-polar tuning capacitor of the ferroresonant core power transformer used in this supply. The two TO-200-case devices in the heat sinks are the 7812 regulator and drive transistor. The two power resistors (lower right) are in the input load-division network of the pass transistors and regulator chip. The two cylindrical objects at the top are tubular capacitors used to bypass AC lines.



Side view of power supply. At left is the 50-amp fuse mounted on two polystyrene pillars. A small bleeder resistor is mounted on the left-hand pillar. The five power resistors on the terminal board comprise the current-sensing resistor ahead of the power transistors. The calibration pot for the voltmeter is mounted on a small tab of circuit board atop the resistor terminal board. The filter capacitors are obvious. At right and behind the large capacitors is the rectifier heat sink with two 35-A 200-PIV diodes for the full-wave center-tap circuit.

the final product

The complete schematic of our finished example power supply is shown in **fig. 9**.

The capacitors in the transformer primary, C1, C2, and C3, are standard equipment and should be included in every power supply. They not only bypass RF that may be present on the power line but also tend to suppress voltage transient spikes coming down the power line by reducing their amplitude before they reach the transformer.

The low-voltage fuse, F2, is placed ahead of the regulator so that it will not degrade voltage regulation. The fuse has appreciable resistance; if it were placed in the output line, the voltage drop would be significant and would vary with the load current, thus degrading regulation.

R1 is simply a bleeder resistor designed to discharge the filter capacitor, C4, when the supply is turned off.

The power supply shown in the pictures is similar, but not identical, to the example supply described in this article. It is built around a 50-ampere saturable-core (constant voltage) transformer and uses six 2N3055 pass transistors in parallel. Four of them are mounted on the long heat sink on top of the supply. Two more are mounted on a smaller heat sink mounted at the back of the supply.

The voltage control pot, R8, is a 10-turn unit with a counter built into the knob, the knob farthest to the right on the front panel of the supply.

A small LED shines through a hole in the panel between the two meters to indicate that power is on.

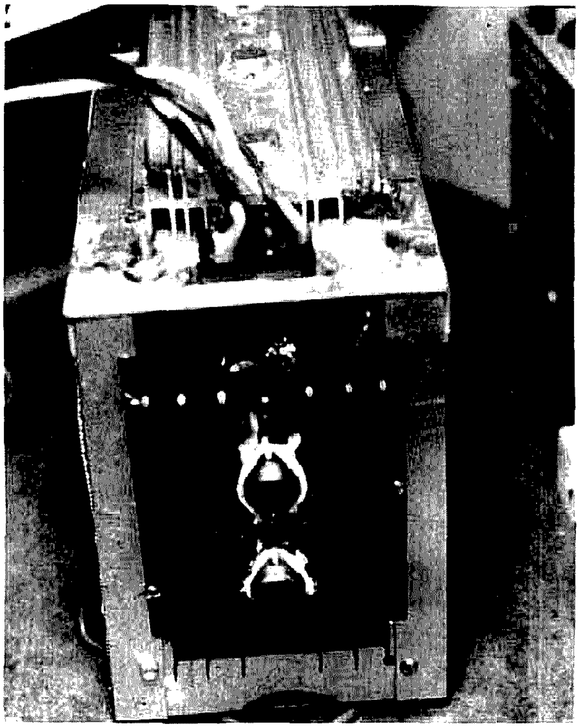
It takes its voltage *from the bridge*, so that it won't be turned on by the battery system that normally floats across the output of the supply.

The sides and bottom, normally enclosed by cane-metal covers, have been left uncovered for photographic purposes. The project was built on a standard light-duty aluminum chassis reinforced with scrap sheet aluminum to support the 30-pound transformer. Pieces of 1-inch aluminum angle stock, mounted on the corners of the chassis, support angle-stock rails at the top. The rails, in turn, support the heat sink.

The 35-ampere, 200-volt rectifier diodes are mounted on a large heat sink that is enclosed by the cane-metal sides. The regulator chip and TIP-125 driver/inverter chip are mounted under the chassis on small heat sinks. All heat-sinked components are insulated with mica washers and silicone heat sink compound.

With the transformer and components shown, the supply can easily deliver 40 amperes or more continuously. The maximum capability of the supply has never been tested, but the pass transistors can easily handle 50 amperes continuously. The transformer was salvaged from a discarded computer terminal unit and the current-handling capacity was estimated from the core cross-sectional area the No. 4 wire used for the secondary center tap serving two secondary windings, each brought out on No. 10 wire. It is a ferro-resonant, saturable core unit with inherent voltage regulation properties providing better than 5 percent load regulation at the bridge output.*

The white pillar to the left of R2 is an insulating post supporting cables going to the B + output connector. The power resistor attached to the post is the bleeder resistor. The chassis-mounted power resistor between the post and R2 is R4. It doesn't need to be that big, but it was the only unit I had with the right resistance value.



A rear view of power supply, with the two pass-transistor heat sinks in view. The output connector and cable appear at the top of the unit. The cable consists of two conductors, each made of two No. 10 stranded wires in parallel.

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19. Budd Meyer, K2PMA, "Low-Cost All-Mode-Protected Power Supply," *ham radio*, October, 1977, page 74.

*The transformer I used is an uncommon and expansive type salvaged from a computer terminal. Suitable transformers of conventional design ranging from $45 to $65, depending on size and features selected, may be obtained from Avatar Magnetics, 1147 N. Emerson Street, Indianpolis, Indiana 46219. Avatar is operated by Ron Williams, W9JVF. Information sheets are free on request.

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ham radio

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fast-scan ATV power amplifier

Modification of 4CX250B design boosts power, improves image

After running QRP (6 watts) Fast-Scan Amateur TV for a year, I decided I wanted to build an amplifier for ATV that would give me a substantial amount of power as well as good quality video, color, and subcarrier sound. After talking with Ron Stefanskie, W9ZIH, who runs *really* high power (500 watts from a commercial broadcast tube — an 8938), I decided I couldn't afford this approach and probably wasn't technically competent enough to build it anyway. Bill Bryant, K9KKL, my mentor, made up my mind for me. After seeing his single 4CX250B run 70 video watts, I started looking for information and parts to build a two-tube 4CX250B (K2RIW design) amplifier. The hardest part was collecting the parts. It took a whole summer of hamfests and a winter of contacting friends to finally gather everything needed.

modifying the linear

After some research, I decided to build the basic K2RIW design. Since I had no plans to run it on SSB, and only on ATV, I built everything heavy duty to withstand continuous-duty video transmissions, and to grid-modulate the tubes with video. Because of the bandwidth restrictions of the grid circuit, the sync pulses, the 3.579 MHz colorburst, and the 4.5 MHz sound subcarrier signals are severely attenuated if you run the amplifier as a linear. But by driving the amplifier with an FM carrier, feeding your video and subcarrier audio via a video modulator into the control grids of the 4CX250B's, you are able to avoid the effects of the grid circuit. Now the only circuit left to affect the video is the plate line, and its bandwidth is wide enough to pass good color, sound, and excellent resolution if it's properly tuned.

Basically, I followed the original construction article for the amplifier with some minor changes.[1] I trimmed the plate stripline inductor, L1, to resonate at 439.250 instead of 432 MHz. The grid feed-through capacitor, originally 1000 pF, was changed to a value that will pass video; I used 60 pF. The screen grids must be video-bypassed with a 10 microfarad capacitor (450 volt rating) as well as RF bypassed with a 1000 pF feed-through. The grid metering cannot be utilized because of the video modulated negative bias voltage. Bypassing is very important for the grid compartment because of high RF fields and the inclusion of video.

Even though the original article called for SK-610 sockets, forget it! Use the SK-620s or SK-630s with the built-in screen-bypass ring. And while the original article called for beryllium copper for the flappers, I used silver plated brass. The plate line and plate load capacitors, as well as the grid line inductors are all silver plated, too.

ready-made modules simplify construction

I used the VM-2 grid modulator and FMA-5 subcarrier audio board available from P.C. Electronics* to modulate the grids. I found several things necessary for good results. Use short shielded leads between the subcarrier audio board and the modulator, and between the modulator and grid feed-through capacitor. All power supplies, modulator supply, grid bias and screen supply should be well regulated. The bias voltage should be adjustable so it can be set for best results. The carrier injection level from the subcarrier audio generator is important, because too high a level destroys the color, and too low a level will eliminate the audio.

The frequency of the audio board should be within ±10 kHz of 4.500 MHz. A series trap consisting of a capacitor and inductor resonant at 4.5 MHz must be placed at the point where the video is fed into the modulator. This keeps the video line from loading down the 4.5 MHz signal.

Probably the most important part of this whole operation is the tuning of the amplifier for a good high resolution and good color with subcarrier audio signal. Tune to the high sideband side of the signal so as not to attenuate the sync pulses and 4.5 MHz audio signal.

---

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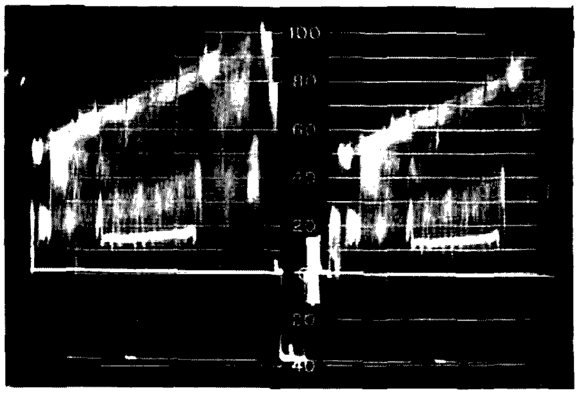
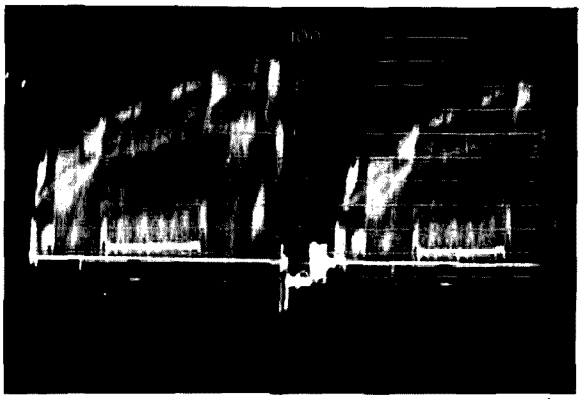
tune-up is easy

To tune the amplifier properly, I strongly recommend using an RF/video detector at the amplifier output. I use a DM-1 RF/video detector board (also available from P.C. Electronics) and use capacitive disc coupling in the plate compartment to sample the outgoing signal and feed it into my detector. This then goes into an RCA TO-1 waveform monitor, although any oscilloscope with a bandwidth of 5 MHz will work. You can observe your transmitted video waveform and the effect tuning the amplifier has on it.

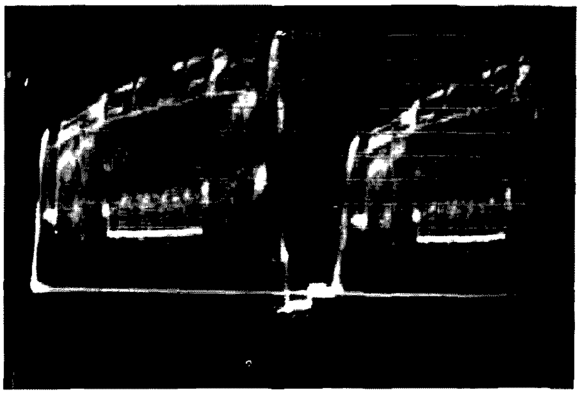
You cannot tune for maximum power out as in SSB. Instead, tune for as much power out as possible while still maintaining a good quality video waveform without clipping sync pulses, color-burst, and 4.5 MHz audio subcarrier signals. I drive my amplifier with an old T-44 Motorola FM transmitter strip with about 10 watts. That's not saying *you* couldn't build up a solid-state 439.250 MHz FM exciter or even drive it with one of the all mode or FM transceivers on the market that covers that frequency. With 10 watts of drive 1350V on the plates, and 400 milliamperes of current at −60



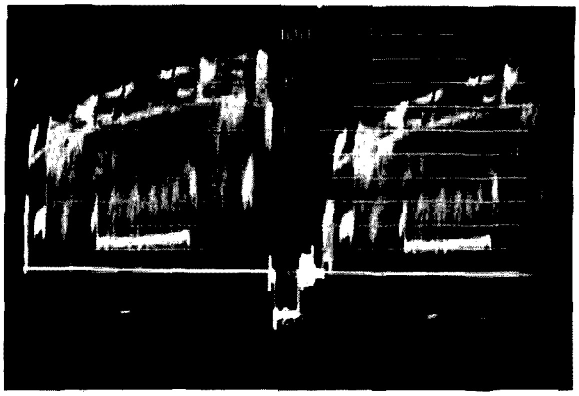
A correct video waveform showing the 0 line or blanking level, the sync pulse, color burst and video information from 0 to 100, 0 being reference black and 100 being reference white.



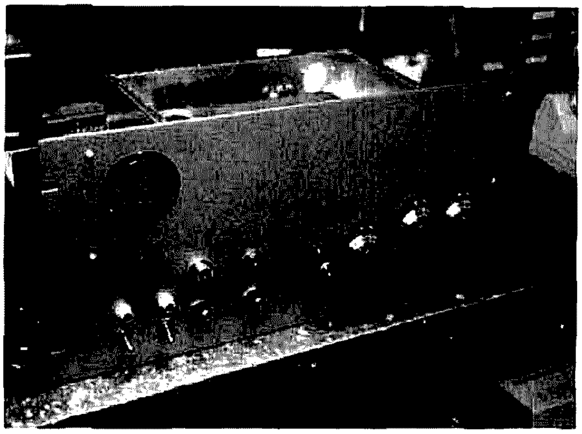
Shown here: the amplifier was tuned for maximum power out, but look at the sync pulse!



Believe it or not, this produced a picture. Shown here is what happens when the amplifier is tuned on the wrong sideband; notice that the sync pulse and color burst signals are gone. (The overall quality of this picture was poor.)



Sync compression is shown here from trying to tune the amplifier for that extra watt of power. What you think you gain in power you lose in the ability for your picture to lock up under weak signal condition.



Faceplate of the amplifier which is rack mounted with the power supply. Notice the "video in" connector (SO-239) and video level adjustment pot and "audio in" (1/4-inch phone plug) and level adjustment pot.

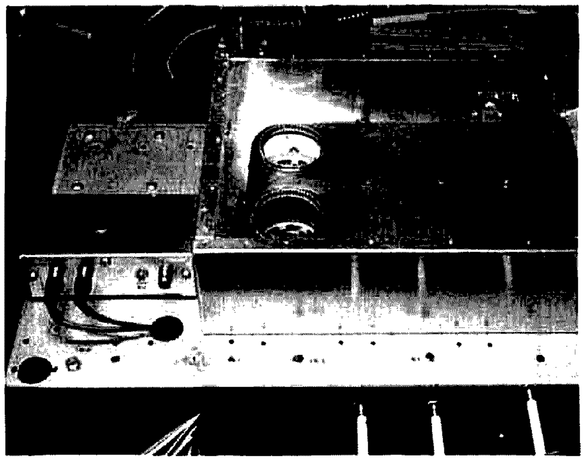
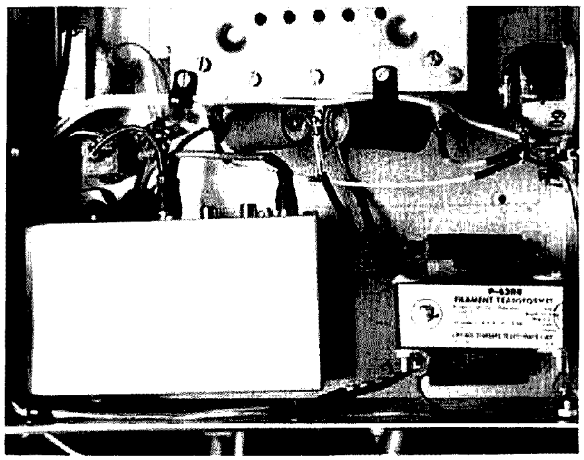


Plate line inductor and compartment, with audio subcarrier generator (left). The 4.5 MHz audio signal goes through the chassis via shielded cables to the modulator located underneath with the grid box.



This photo shows the grid modulator (top) in relation to the grid compartment. Notice 10 microfarad, 450 volt bypass capacitors on the grid box for the screen. Filament transformer is shown in the lower right; the modulator should be close to grid feed-through and shielded cables used.

volts of bias, I get 200 watts of acceptable quality color video and sound. For black and white operation only and without sound, I can push it to over 300 watts. Run properly, the hottest air temperature from the tubes is 115 degrees, and this is after one-hour of key down! So for those of you who have an existing RIW type amplifier or are thinking of building one to operate on 439.250 MHz Fast-Scan ATV, for a little extra effort and a few extra parts, you can be rewarded with high power, high-resolution FSATV pictures.

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ham radio

predicting solar outages

Don't worry. We're not in imminent danger of losing the sun. The term "solar outage" merely describes the phenomenon in which radio reception suddenly deteriorates, with a decreasing signal-to-noise ratio. This article is addressed to Radio Amateurs experimenting with the reception of signals from the geosynchronous satellites operating in the 3.7-4.2 GHz region (not an Amateur band). — Ed.

A solar outage occurs when a particular satellite, as "seen" by the antenna of a receiving station on Earth, is directly in line with the sun. This phenomenon occurs only twice a year for only a short period of time, with the duration of the outage primarily determined by the size of the antenna and its elevation.

Used in conjunction with the information provided in **table 1**, the computer program provided in **fig. 1** allows the determination of the date of solar outages. Declination of the sun, a necessary input, can be found in any current almanac or calculated using **eq. 1**.

Alternatively, the following formula can be used to calculate the sun's declination.

$$Dec = ARCSIN\ (SIN_{El} \times SIN_{LAT} + COS_{El} \times COS_{AZ} \times COS_{LAT}) \quad (1)$$

where Dec = sun's declination (measured in degrees)
El = elevation angle (from the antenna to the satellite)
Lat = latitude of the receiving station (measured in degrees)
Az = azimuth or bearing from the antenna to the satellite (measured in degrees)

Use either method to determine the value of declination needed for insertion in the program listed in fig. 1.

During my first year of TVRO experimentation I built a parabolic dish faced on the inside with tinfoil. In my haste to put it into operation in time for Christmas, I postponed painting the foil until early spring, when the weather would be a little warmer. On February 26th, when I arrived home late in the afternoon, I noticed that part of the feedhorn was lying on the ground, looking as if it had been shot. Though the temperature was in the 30's, the focusing action of the dish produced temperatures *over 2000 degrees F (933 degrees C).* Luckily the LNA was not damaged.

This shouldn't happen with today's dishes, although *I have seen a number of unpainted dishes in use.* Remember to be very careful with that huge solar mass; I've since learned to respect the sun.

The program shown in **fig. 1** is written for the Radio Shack TRS-80 color with extended BASIC. With minimal changes, the routine should work with most other personal computers as well. Remember that most computers work in radians rather than in degrees, so degree value must be divided by (180/pi). At the end of the program, multiply the value by 57.295 to obtain degrees.

provide these inputs:

EL — See any satellite tracker program. (One is available from the author. See details at end of article. — Ed)
AZ — See above
LAT — Check with local government agencies or purchase 7.5 minute topographic maps from local stationary or book stores.

Example. The elevation of the antenna is 31 degrees and azimuth is 197 degrees. The antenna latitude is

By Vern Epp, VE7ABK, 705 6th Street, Nelson, B.C., Canada

```
5 REM/BY VERN EPP 705 6TH ST.,
7 REM/NELSON B.C. CANADA
8 REM/COPYRIGHT (C) 1984
10 CLS
20 PRINT"*****SATELLITE SOLAR OUTAGES****"
30 PRINT:INPUT"LAT. IN DEGREES IS";D
40 INPUT"LAT. IN MINUTES IS";M
50 INPUT"LAT. IN SECONDS IS";S
60 LAT=D+M/60+S/3600
70 INPUT"ELEVATION IN DEGREES IS";EL
80 INPUT"AZIMUTH OF SATELLITE DISH IN    DEGREES IS";AZ
90 PI=3.1416
100 DR=180/PI
110 DEC=(SIN(EL/DR)*SIN(LAT/DR)+ COS(EL/DR)*COS(AZ/DR)*COS(LAT/D
R))
120 X=ATN(DEC/SQR(DEC*DEC+1))*57.295
130 PRINT"THE TRUE DEC IS ";:PRINT USING"###.###";X;:PRINT" DEGR
EES"
140 PRINT@462,"WANT ANOTHER?"
150 A$=INKEY$:IF A$=""THEN 150
160 IF A$="Y" THEN 10
```

fig. 1. TRS-80 color computer program listing helps determine the sun's declination angle throughout the year.

table 1. General-purpose almanacs provide declination values in minutes and degrees; for use in program, these must be converted to decimals.

Feb		15	-02.25	09	+07.47	Sept		27	-01.52
19	-11.37	16	-01.85	10	+07.85	01	+08.38	28	-01.92
20	-11.02	17	-01.45	11	+08.22	02	+08.02	29	-02.30
21	-10.65	18	-01.07	12	+08.58	03	+07.65	30	-02.68
22	-10.30	19	-00.77	13	+08.95	04	+07.28		
23	-09.93	20	-00.27	14	+09.32	05	+06.92	Oct	
24	-09.57	21	+00.13	15	+09.67	06	+06.55	01	-03.08
25	-09.18	22	+00.52	16	+10.03	07	+06.17	02	-03.47
26	-08.82	23	+00.92	17	+10.38	08	+05.80	03	-03.85
27	-08.45	24	+01.32	18	+10.73	09	+05.42	04	-04.23
28	-08.07	25	+01.70	19	+11.08	10	+05.05	05	-04.63
		26	+02.10	20	+11.42	11	+04.67	06	-05.02
March		27	+02.48			12	+04.28	07	-05.40
01	-07.68	28	+02.88			13	+03.90	08	-05.78
02	-07.30	29	+03.27	Aug		14	+03.52	09	-06.15
03	-06.92	30	+03.67	20	+12.53	15	+03.13	10	-06.53
04	-06.53	31	+04.05	21	+12.22	16	+02.75	11	-06.92
05	-06.15			22	+11.88	17	+02.37	12	-07.30
06	-05.77	April		23	+11.53	18	+01.98	13	-07.67
07	-05.38	01	+04.43	24	+11.20	19	+01.58	14	-08.05
08	-05.00	02	+04.82	25	+10.85	20	+01.20	15	-08.42
09	-04.60	03	+05.20	26	+10.52	21	+00.82	16	-08.78
10	-04.22	04	+05.58	27	+10.17	22	+00.42	17	-09.15
11	-03.82	05	+05.97	28	+09.82	23	+00.03	18	-09.52
12	-03.43	06	+06.35	29	+09.45	24	-00.35	19	-09.88
13	-03.03	07	+06.73	30	+09.10	25	-00.75	20	-10.23
14	-02.63	08	+07.10	31	+08.75	26	-01.13	21	-10.60

table 2. Number of solar outage days and duration.

antenna diameter (feet)	8	10	12	16	20
outage	16	14	12	10	8
outage during these days (minutes)	30	24	20	15	12

Note: To determine the beginning of the outage period find the *middle* day (using **fig. 1**) then use this table to locate, under your antenna size, the total span of outage days and divide by 2.

49 degrees. These values, when entered, provide a declination of −8.37. The almanac indicates these dates as February 27 and October 15th. From the outage days chart (**table 2**) we find that by using a 10-foot dish, we have solar degradation for ±7 days for a period of approximately 24 minutes.

The outage chart shown in **table 2** is designed for 4 GHz operation and provides the total number of days in which some solar degradation can be expected to occur. The time, expressed in minutes, represents the approximate duration of the anticipated outage.

Information provided in the appendix can be used to determine the exact time of day which the outage will occur.

reference

1. Dennis Mitchell, K8UR, "Receiving Signals from Space," *ham radio*, November, 1984, page 37.

appendix

time of the day calculations

Step 1

$$LHA = ARC\,COS\left[\frac{SIN_{EL} - (SIN_{LAT} \times SIN_{DEC})}{COS_{LAT} \times COS_{DEC}}\right]$$

Step 2

$$GMT \cong \frac{(LHA + longitude)}{15} - 12$$

LHA Local angle hour is measurement of the sun's current position (measurement in degrees)
EL Elevation of dish on earth
LAT Latitude in degree of earth station
DEC Declination of the sun measured in degrees
GMT Greenwich Mean Time
LONGITUDE Earth station in degrees

Note: GMT must be converted to local area time

Vern Epp's computer program, "Satellite Tracker, ($34.95) includes routines such as geo-sat location, dish design and efficiency evaluation, polar Dish Design LNA conversions, and C/N calculation. Also available is a program for predicting solar outages, ($11.95) that calculates outage times using built-in charts. A number of other useful routines are included. Both programs are written for TRS-80C extended BASIC computer. For details, contact the author at the address given on page 75.

ham radio TECHNIQUES

Bill Orr, W6SAI

It may be too early in the year to start working on your antenna, but it's not too soon to start thinking about interesting antenna designs! Here's a good one for you.

the "K3ZVH Special" for 40 through 10 meters

This antenna design came to me via Jack Walker, VE4DS, who got the information from J.A. Hutcheson, K3ZVH, who has a summer home in Selkirk, Manitoba, where Jack lives. The antenna covers 40, 30, 20, 15, and 10 meters, and provides a low value of SWR on each band (**fig. 1**). It is a "fat," center-fed, broadband dipole. Old Timers may find a tear in their eyes as it brings back memories of the old cage antenna, so popular in the early 1920's. This design, however, works on a different principle.

The overall length of the antenna is 85 feet (25.9 meters) and cage diameter is 2 feet (0.61 meter). The antenna is fed at the center with a 300-ohm, two-wire line. VE4DS uses open-wire "ladder" line and K3ZVH uses good-quality 300-ohm transmitting ribbon line. Line length is not critical. A 4-to-1 balun provides a good match to a 50-ohm coaxial line.

building the antenna

The antenna is built of No. 16 enamel wire; the hoops are made of aluminum TV ground wire bent into a circle. The ends of the hoops are joined by a short piece of copper tubing, which is used as a compression sleeve. The wires are attached to the hoops by small pieces of copper wire twisted around the joint.

The assembly can be quite unwieldy unless it is strung out in a line under tension, preferably slung between two low supports. (It would be logical to string one wire of the cages between the supports and then build the rest of the cage around that wire.) The wires are spaced equidistant around the loops, and the tie-wires are wrapped with electrical tape after the assembly is finished. Because it doesn't seem to matter whether electrical contact is made between the loops and the antenna wires, no effort was made to make such a connection.

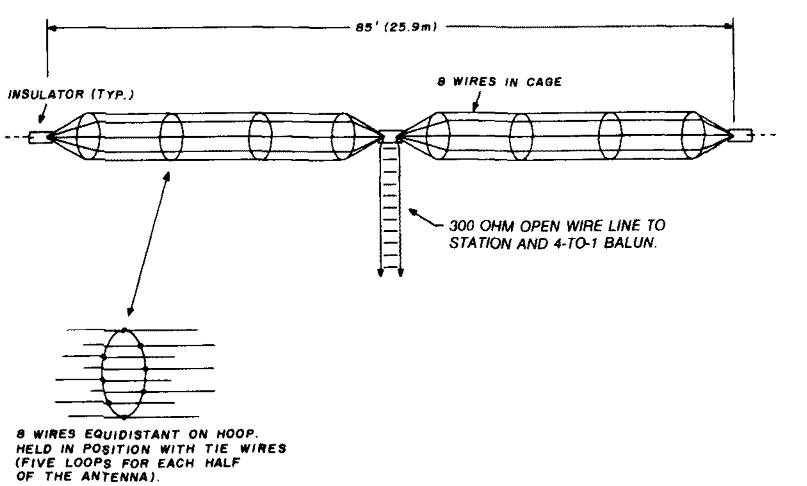
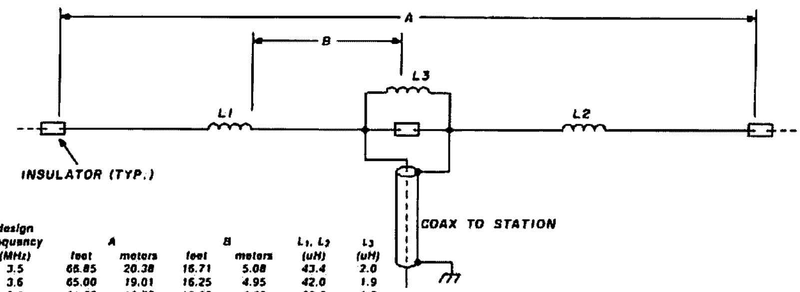


fig. 1. The K3ZVH "fat dipole" for 40 through 10 meters.

For lowest SWR (usually required with a solid-state transmitter), a simple antenna matching unit is placed at the transmitter. If the transmitter can work into an SWR as high as 2:1, no tuner is required.

some thoughts on the K3ZVH antenna

It would be interesting to run an SWR plot of this antenna over the spectrum from 5 MHz to 30 MHz. Because of the "fat" design, I would think that impedance excursions from a mean value would be modest and the antenna might qualify for operation over the whole frequency range. Many military "broadband" antennas are built in this fashion. If any reader of



design frequency (MHz)	A feet	A meters	B feet	B meters	L_1, L_2 (uH)	L_3 (uH)
3.5	66.85	20.38	16.71	5.08	43.4	2.0
3.6	65.00	19.01	16.25	4.95	42.0	1.9
3.8	61.57	18.77	15.39	4.69	39.6	1.8
4.0	58.50	17.83	14.62	4.46	37.3	1.7

antenna wire size: 14 AWG
L_1, L_2, L_3 diameter = 2 inches (5 cm)
(all coils are close wound)

3.5 L_1,L_2 = 36.5 turns, L_3 = 4.8 turns
3.0 L_1,L_2 = 35.0 turns, L_3 = 4.7 turns
3.8 L_1,L_2 = 34.0 turns, L_3 = 4.6 turns
4.0 L_1,L_2 = 32.5 turns, L_3 = 4.4 turns

fig. 2. Compact dipole design for 80 meters.

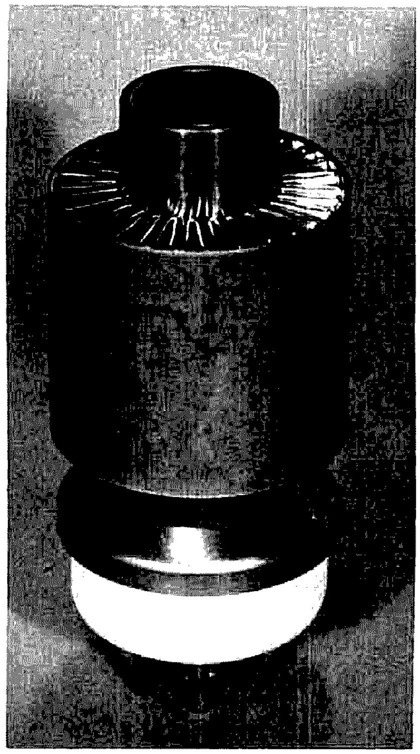


fig. 3. The new EIMAC 3CX1200A7 high-μ power triode.

table 1. Characteristics and operating parameters of the EIMAC 3CX1200A7.

plate voltage	3600 volts
cathode voltage	3.4 volts
zero-signal plate current	220 mA
single tone plate current	700 mA
two-tone plate current	500 mA
single-tone grid current	215 mA
two-tone grid current	130 mA
useful power output (CW or PEP)	1750 watts
3rd-order IMD products	−34 dB
single tone drive power	108 watts
filament voltage	7.5 VAC
filament current	21 amperes
input capacitance	20 pF
output capacitance	0.2 pF

cooling: 30 CFM at 0.5 inch/water (anode to base)
(cooling air must be supplied to base)

Note: Zener bias derived from four G.E. 1N5062 diodes in series, forward biased, or Motorola 1N4549.

this columh would care to make a frequency-sweep of this design, I'd be happy to hear about the results.

the 80-meter compact dipole

In my February column I described a half-size 160-meter dipole for those unfortunate hams who do not live on 200 acres atop a high hill. The same technique can be used to make a compact 80-meter dipole that will give a good account of itself (**fig. 2**). This dipole is only about 65 feet (19.8 meters) long and has a bandwidth of about 50 kHz where the VSWR is less than 2:1. Antenna length is a compromise between efficiency and bandwidth and the same truths that apply to the 160-meter antenna shown last month apply to this antenna: a loaded antenna can be just about any length, but the tradeoffs must be accepted. This design is a good compromise, and I recommend it.

Three coils are needed — two loading coils to be placed in the flattop and an impedance matching coil placed across the feedpoint. A balun is not required and the antenna is fed directly from a 50-ohm coax line.

adjusting the antenna

Once the antenna has been built, it is erected in place, but at a low elevation so that a dip-meter can be coupled to the matching coil, L3. The end sections are trimmed equally until resonance is established for low-end (3.5-3.7 MHz) or high-end (3.8-4 MHz) operation. The feedline is removed for this adjustment.

When antenna resonance is established, the antenna is pulled up to its final position and SWR measurements are made across the frequency band of operation. Matching coil L3 is adjusted, a quarter-turn at a time, for the lowest SWR on the feedline at the frequency of antenna resonance. Refer to the discussion of the 160-meter antenna in my February column for details of the loading and matching coil construction and installation.

the new EIMAC 3CX1200A7 power triode

In recent months Varian EIMAC has released data on a new tube of interest to Radio Amateurs — the 3CX1200A7 (**fig. 3**). For Old Timers, I can say it is the "Son of 3-1000Z." For newcomers, it is a 1200-watt dissipation,

high-μ power triode intended for cathode driven, linear amplifier service. The tube is very compact, being only 6 inches (15.3 cm) high and 2-5/8 inches (6.7 cm) in diameter. The tube is air-cooled and is rated for full input to 110 MHz.

The characteristics and typical operating parameters of the 3CX1200A7 are listed in **table 1**.

Using the 3CX1200A7

The 3CX1200A7 is well-suited for linear amplifier service because it provides maximum FCC-rated power output at a drive level compatible with today's modern solid-state exciter.

A representative circuit for the 3CX1200A7 in linear service is shown in **fig. 4**. This is a cathode-driven, grounded-grid configuration. A π-L network output circuit is chosen for maximum harmonic attenuation and a π-network input circuit is used to provide a good match between the exciter and the amplifier (**tables 2** and **3**). Current metering is done in the return leads to the power supply so no high voltage appears on the meter movements.

Standby and operating bias voltages are provided in the grid-filament return circuit. Operating bias is set by a Zener diode to approximately +3.4 volts. The resting plate current is determined by the value of zener bias. With no bias, at a plate potential of about 3600 volts, resting plate current is nearly 250 mA, so a small value of bias lowers the current and reduces plate dissipation. For standby, additional cathode bias is added in the form of a 10k resistor to reduce the current to a very low value. The resistor is shorted out by contacts on the VOX relay.

A 50-ohm wirewound resistor from the negative side of the plate supply to ground makes certain that the negative terminal does not rise to the value of the plate voltage if the positive side of the supply is accidentally shorted to ground.

Two reverse-connected diodes are shunted across the safety resistor to limit any transient surges under a shorted condition that might cause wiring insulation breakdown or meter damage. A resistor across the zener diode provides a constant load for it and prevents cathode voltage from soaring if the zener safety fuse should open.

A 10-ohm, 50-watt resistor is placed in series with the B-plus lead to the plate RF choke. The resistor serves as a low-Q VHF choke to suppress harmonic currents in the power lead and also protects the tube and associated circuitry in case of a flash-over in the tube or plate circuit. The considerable amount of energy stored in the power supply filter capacitor is instantaneously "dumped" into the amplifier in the unlikely event of a flash-over, and much of this destructive energy is dissipated in the resistor.

Parts list, 3CX1200A7 amplifier.

item	*description*
C1,C2	*see table 2*
C3,C4	*see table 3*
CR1	*10 volt, 50 watt zener diode*
J1,J2	*coaxial chassis connector (SO-239 or equivalent)*
J3	*high voltage connector*
M	*0-10 volt iron-vane type AC meter (RMS responding)*
M1	*0-1 ampere DC*
M2	*0-500 mA DC*
PC	*three 100-ohm, 2-watt resistors in parallel shunted by 3 turns No. 14 AWG, 0.5-inch diameter, 0.75-inch long. Coil may be wound around one resistor.*
RFC1	*50 μH. 14 bifilar turns No. 10 AWG enamelled wire wound on ferrite core 5 inches long, 0.5-inch diameter (Indiana General CF503 or equivalent)*
RFC2	*100 μH, 1 ampere DC; 112 turns No. 26 AWG space wound on 1-inch ceramic or teflon form 6 inches long. Series-resonant at 24.5 MHz with terminals shorted (B&W 800A, or equivalent)*
T1	*7.5 volts at 21 amperes, tapped primary*
TD	*time delay relay (3 seconds) Amperite 115-NO3*

Note: All capacitors other than C1 through C4 are 0.01 μF, 600 volt. Capacitors are mica type.

table 2. Cathode circuit component values with an R_C of 60 ohms.

band	C1 (pF)	C2 (pF)	L1 (μH)
160	3300	3100	3.53
80	1700	1670	1.90
40	900	840	0.96
20	440	417	0.47
15	300	275	0.32
10	220	205	0.23

table 3. Plate circuit component values with an R_L of 2750 ohms.

band	C3 (pF)	C4 (pF)	L2 (μH)	L3 (μH)
160	330	1550	33.0	11.2
80	165	775	16.5	5.6
40	80	385	8.2	2.7
20	40	190	4.1	1.4
15	27	130	2.7	0.9
10	20	100	2.0	0.7

Note: Data on plate circuit design can be found in the 22nd edition of the *Radio Handbook*, available from Ham Radio Bookstore, Greenville, NH 03048.

the step-start circuit

The resistance of the thoriated tungsten filament of the 3CX1200A7, when cold, is about one-tenth its value at operating temperature. As a result,

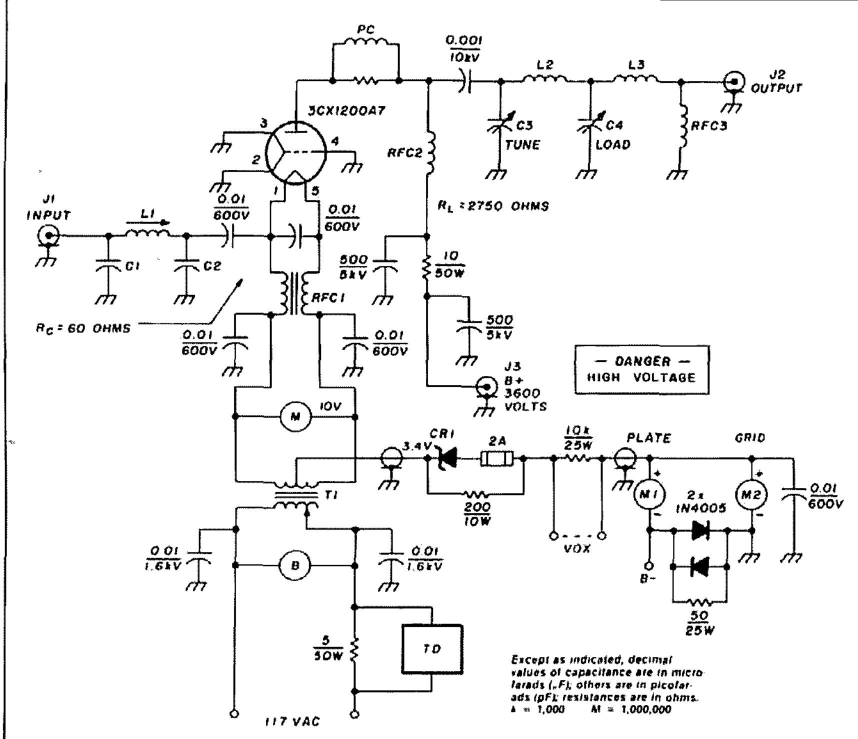


fig. 4. Schematic, high-power amplifier using EIMAC 3CX1200A7 tube.

a filament inrush current as high as 210 amperes may occur for a fraction of a second. It is good commercial practice to limit filament inrush current as this powerful surge can warp or otherwise strain the filament structure of the tube. This design incorporates a simple inrush time delay (TD) protection circuit that applies reduced voltage to the tube filament for a short period, then allows application of full voltage when the filament has had a chance to warm up a bit. Total time delay is only a matter of seconds, and is a wise precaution because the circuitry costs but a fraction of a replacement tube!

safety factors

It's a good idea to use a filament transformer having a primary winding tapped for various line voltages. A filament voltmeter can provide the operator with a close check on voltage, which should be held to ±5 percent of the nominal value of 7.5 volts.

Complete, detailed information on the design and construction of high-power amplifiers of this type can be found in the 22nd edition of the *Radio Handbook*, available from Ham Radio's Bookstore, Greenville, New Hampshire 03048.

radio tube closeout

Edlie Electronics is closing out their surplus tube department. Their 1985 catalog lists over 550 types of receiving tubes at two for a dollar, ten for $3.95, and 50 for $15. All have been tested, but are sold "as is" with no guarantee. Most are boxed.

For details, contact Edlie Electronics, 2700 Hempstead Turnpike, Levittown, New York 11756 (516-735-3330).

list of EME (moonbounce) stations

A complete list of U.S. and overseas VHF moonbounce stations is now available. For a copy, send five first-class postage stamps to me at EIMAC, 301 Industrial Way, San Carlos, California 94070 U.S.A.

ham radio

get on 6 meters — the inexpensive way

Converting the CRT-1/CPRC-26

The CRT-1/CPRC-26 radio set shown in **fig. 1** is an ideal unit for low cost, multichannel 6-meter FM operation. A Canadian-designed and manufactured six-channel, crystal-controlled receiver-transmitter, it covers the 47.0 to 55.4 MHz frequency range with a power output of 300 milliwatts. The receiver's sensitivity is 2 microvolts; its deviation is approximately 15 kHz. All that's necessary to make the CRT-1/CPRC-26 operational is a power supply and the Canadian version of the U.S. Military H-33/PT Handset. You can even modify the basic unit described in this article to include speech limiting, squelch circuitry, and provisions for loudspeaker operation.

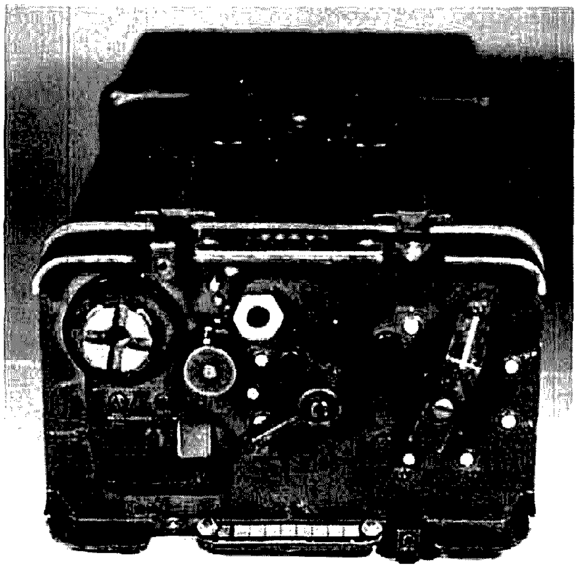


fig. 1. The CRT-1/CPRC-26 radio set represents an easy way of "getting on" 6 meters.

theory of operation

Figure 2 shows a block diagram of the radio set. Incoming signals are amplified in the RF amplifier, V7, and applied to the mixer, V8, along with the output of the crystal oscillator, V4, to produce an IF of 4.3 MHz. The crystal frequency equals the frequency of operation minus 4.3 MHz. The IF signal is then amplified through four identical IF amplifier stages and fed to the limiter, V10. The signal is then routed through the audio discriminator to the audio amplifier, V6, and to the earphone element of the handset via the output transformer, T1.

Audio input from the microphone element of the handset is applied through the microphone transformer, T2, to the modulator, V3. The amplified audio is next fed to the master oscillator, V2, via the master oscillator coil. The audio input varies the frequency of the master oscillator to produce direct FM. The output from the master oscillator, V2, is then applied to the power amplifier, V1, and routed to the antenna via the output tuning network.

Because the master oscillator is essentially a VFO, some method is needed to ensure that the transmitter frequency tracks the receiver frequency. Therefore, the output from the mixer, V8, is fed to the AFC amplifier, V5, during transmit, through the AFC discriminator to the master oscillator, V2, via voltage dividers R5 and R6. In effect, output from the crystal oscillator, V4, provides transmitter frequency control.

power supply

The power supply for the radio set consists of an inverter for high and low B + and a simple + 1.4-volt

By **Mark Starin, KB1KJ,** 457 Varney Street, Manchester, New Hampshire 03102

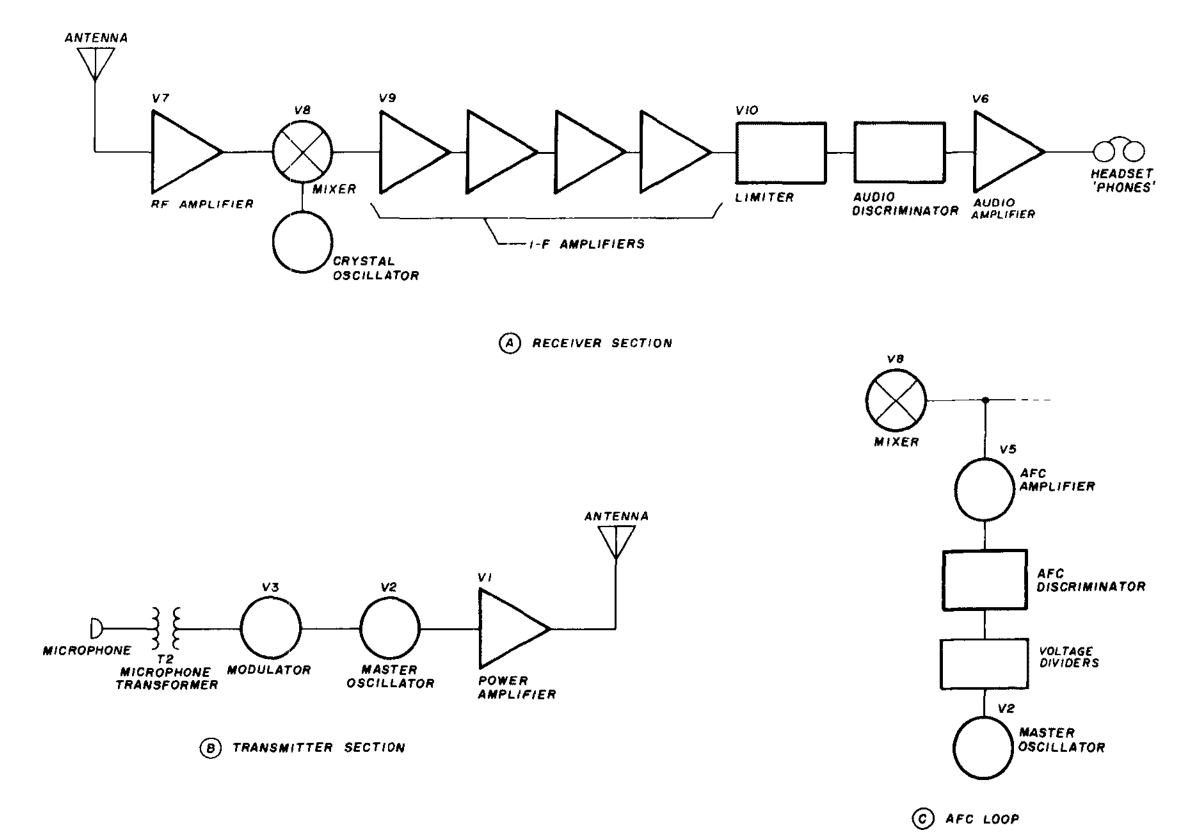


fig. 2. The CRT-1/CPRC-26 radio set block diagram (*A*) receiver section, (*B*) transmitter section, and (*C*) AFC loop.

filament regulator board. Bias is supplied by two +1.5-volt N cells connected in series and mounted on the filament regulator board. Both the inverter and the filament regulator board fit inside the battery box, which attaches to the rear of the radio set. A schematic of the filament regulator circuit is shown in fig. 3.

The inverter can be either bought or built. A surplus transistorized inverter module, P/N 522-1091-004, is available from Fair Radio Sales Company, Inc., (see fig. 4). The only disadvantage of this choice is that unless the inverter is modified, a +24-volt DC power source will be needed. This might prove to be a problem for anyone operating from a typical automobile. A schematic diagram for this circuit is shown in fig. 5. With +13.8 volts power applied to U1, the input regulator circuit, R1, is adjusted until the output voltage present at C3 is +140 volts maximum. This occurs when Q1 and Q2 are biased into oscillation by R3 and R4, which produces high-voltage AC at the secondary of transformer T1. This AC voltage is rectified by CR1, filtered by C3, and applied to the input of the U2 output regulator circuit. Note that the LM317 is capable of high-voltage regulation provided the input voltage does not exceed the output voltage

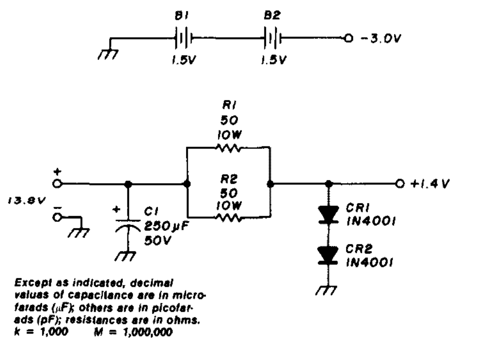


fig. 3. The filament regulator circuit uses a pair of 1N4001s to meet the 1.4 volt bias requirements.

by more than 40 volts. Therefore, I recommend that the input regulator be set to produce approximately 130 volts at the input of U2 (approximately +5 volts to the transistors). CR2 is a 40-volt zener diode that protects U2 if the input regulator circuit fails. Note that a simple voltage divider, consisting of R6 and R7, provides approximately +45 volts for the receiver and transmitter.

The +1.4 volt filament regulator board consists of a series dropping resistor (two 50-ohm, 10-watt resistors in parallel), two 1N4001 diodes shunted to ground, and a filter capacitor. Although there are alternatives that will provide +1.4 volts for the filaments, this circuit is especially simple.

The two N cells are another way to provide needed voltage — in this case, the bias voltage for the receiver audio amplifier. The N cells are mounted inside a standard 9-volt battery holder and secured to the battery holder with nylon ties. The battery holder is then mounted on the filament regulator board. Transistors Q1 and Q2 are mounted on the top of the battery box. Their placement is not critical; almost any convenient location on the battery box will probably work satisfactorily. The transformer is mounted inside the battery box on either the left or right side. The filament regulator board is also mounted inside the battery box on the side opposite the transformer. Connections from the power supply to the radio set are made with hookup wire wrapped around and soldered to the pins on the rear panel of the radio set. Heat-shrink tubing is then placed over these connections. In addition, a power cable was fabricated using a Radio Shack two-pin plug and matching jack (P/N 274-201 and 274-202), 12-gauge wire, and battery clips (Radio Shack P/N 270-344). **Figure 6** shows the power supply mounted in the battery box with the power cable attached. The dimensions of the filament regulator board are 1 inch

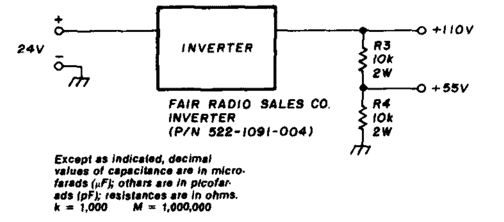


fig. 4. The inverter module, available from Fair Radio Sales, requires a +24 volt source of power.

table 1. Alignment procedure.

VTVM	channel	adjust	indication
2-volt scale, −DC position (Insert the probe into pin 3 of the test socket.)	1-6	C3-C8 C29-C34	maximum −1 volt
10-volt scale, +DC position (Insert probe into pin 4 of test socket. Press PTT (push-to-talk) switch on handset.)	1-6	C19-C24	zero indication between positive and negative peaks

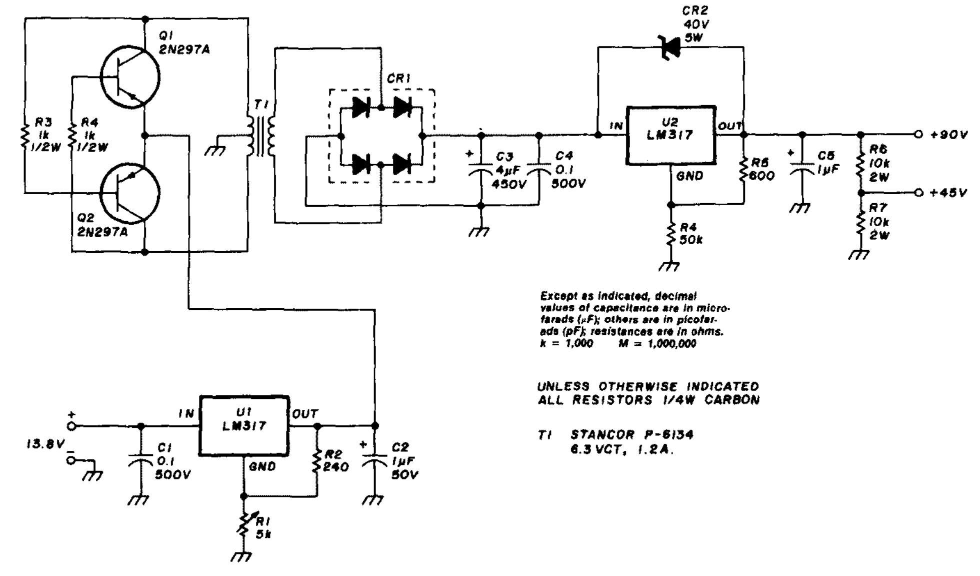


fig. 5. An inverter can be easily built around two LM317 ICs.

fig. 6. The complete radio set, shown with power supply and cable, measures 3 × 4.75 × 10.34 inches (7.62 × 12.07 × 26.26 cm).

(25.4 mm) by 2 inches (50.8 mm). The set's power requirements are as follows:

receiver:	+ 1.5 volts, 550 mA;
	+45 volts, 12 mA;
	+90 volts, 30 mA;
	−3 volts bias
transmitter:	+ 1.5 volts, 850 mA;
	+45 volts, 8 mA;
	+90 volts, 30 mA.

alignment

The alignment procedure for the radio set is easy and does not require elaborate test equipment. A VTVM and a 5/32-inch color TV tuning wand are necessary.

First remove the battery box from the rear of the radio set. Then remove the cover from the receiver-transmitter chassis. Apply +90, +45, + 1.5, and −3.0 volts to the radio set. (Don't forget to connect a dummy load to the antenna connector on the front panel.) Install the appropriate crystals for the desired operating frequencies. Set the OFF/QUIET/LOUD switch to either the QUIET or LOUD position. **Figure 7** shows the radio set adjustment locations. **Figure 8** shows the major component locations. **Table 1** identifies the controls and proper indications to be observed during the alignment procedure.

obtaining materials

The primary source of the CRT-1/CPRC-26 Radio Set is Fair Radio Sales Co., 1016 East Eureka, Box 1105, Lima, Ohio 45802-1105. Their 1984 catalog price for the radio set — described as "used" — is $12.95. (The "used" radio set we bought appeared to be in

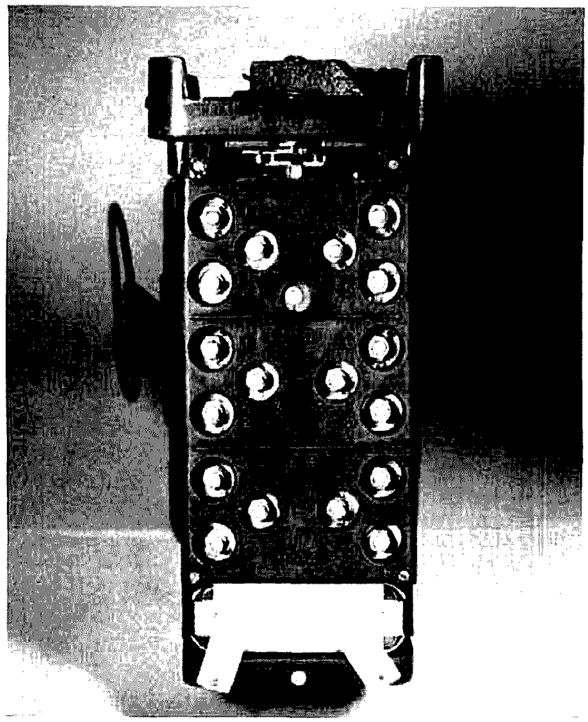


fig. 7. Alignment proceeds smoothly with easy access to the adjustable components.

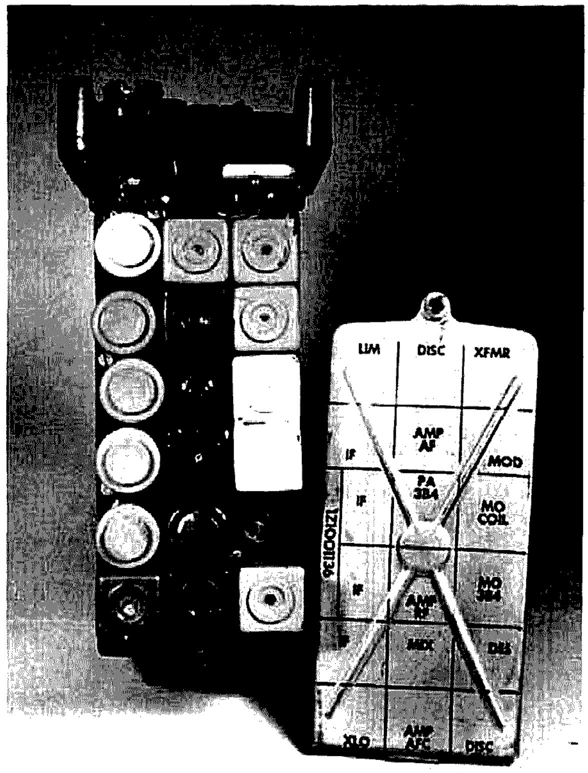


fig. 8. Exposed view of radio set with cover removed shows the location of the major components.

table 2. Spare module color codes and availability.

module name	color	availability (1984)
RF amplifier	black	no
mixer	brown	no
IF amplifier	orange	yes
limiter	yellow	no
audio discriminator	light green	no
audio amplifier	dark green	yes
crystal oscillator	red	yes*
AFC amplifier	dark blue	no
AFC discriminator	light green	no
mo coil	white	no
modulator	grey	yes
transformers	light blue	no

*"RF oscillator" is stamped on the module housing.

like-new condition.) The handset is also available at $8.95 and the schematic at $1.50. Fair Radio also sells spare modules at $1.50 each or 10 for $10 (you mix and match them). The inverter module is available from Fair Radio Sales for $9.95. **Table 2** shows the color codes for the various spare modules and whether they were available from Fair Radio in 1984.

Crystals for the radio set are normally CR-52A/U types that are not currently sold by Fair Radio. We were fortunate to obtain a complete CK-6/PRC-6 quartz crystal unit set for an AN/PRC-6 that saved us the expense of buying individual crystals. (This crystal pack occasionally shows up at flea markets.) However, third overtone crystals (Type EX, available from International Crystal Manufacturing) should also work, although we haven't tried them.

operation

After installing the power supply, connect +13.8 volts to the battery box rear connection, an antenna to the front panel mounted BNC connector, and the handset to the audio connector on the front panel. Set the OFF/QUIET/LOUD switch to either QUIET or LOUD. (Note that there is no deviation or microphone gain control on the radio set.) Because most of your contacts will probably be with stations operating narrowband equipment, back away from the microphone element on the handset when transmitting; this will help to minimize over-deviation.

Any antenna — from a quarter-wave whip to a multi-element Yagi or quad — can be used with this radio set. Although the 300 milliwatt output is definitely in the QRP category, a decent base station or mobile antenna will provide this little rig with a reasonable line-of-sight range.

acknowledgement

The author wishes to express his appreciation to Hal Weinstein, K3HW, Rich Royer, W1HZN, and Gene Balinski, WA1UXA, for their contributions to this article.

ham radio

controlled vertical radiation rhombics, part 1:

designing for high performance

For BIG performance — in a BIG backyard — try rhombics

Rhombics aren't for everyone. Not many Amateurs have sufficient space on their small city or suburban lots to accommodate the generous dimensions of this useful, efficient antenna. Mine measures 277 feet on each leg, and requires an area that measures 525 by 250 feet. Yet for those who have adequate land, rhombics can offer unusually high performance: the three wavelengths on my leg rhombic, for example, puts an *outstanding* signal into Europe on 30 meters, and when the 18 and 24 MHz bands are open, this antenna is all set to go.

In the mid-1930's, when the rhombic antenna was relatively new, ARRL experimenters[1] found it to be useful for long-distance radio communications. League personnel strung hundreds of feet of wire, in an enormous diamond configuration, through the Connecticut woods with astonishing results: signals from as far away as Australia were clearly heard.

As a young man, I duplicated the League's efforts behind my home in New Jersey. After climbing many trees, stringing springy lengths of No. 12 copperweld wire among and through their innumerable branches, I succeeded in constructing a rhombic antenna terminated in a 100-watt carborundum resistor and fed with a 600-ohm open-wire line. It was not until retirement that I was able to build other rhombics; this article, a result of these later experiments, describes the design of a rhombic that operates on the 160 through 10 meter bands and offers controlled vertical radiation in the 40 through 10 meter bands. This feature allows mechanical tuning to the best operating configuration for any band. (Fixed rhombics are best for one frequency and perform well over a 2 to 1 frequency range only.)

Because this rhombic is terminated in its characteristic impedance, it is nonresonant; its input impedance is essentially flat over its entire operating range. In steady use since the 1983 Radiosport contest — when I heard on 15 meters, such reports as "You're the only U.S. station coming through at this time," and "Your signal is overriding all other U.S. stations by 10 to 15 dB" — it has performed admirably and presented no maintenance or durability problems.

how does a rhombic work?

To begin, let's consider the rhombic's horizontal radiation plane, realizing that a rhombic is simply four long-wire antennas arranged in the shape of a rhombus or "diamond." By terminating the rhombic in its characteristic impedance (with a noninductive resistance), a unidirectional lobing pattern is obtained in the direction of the terminated end; (see **fig. 1**). The

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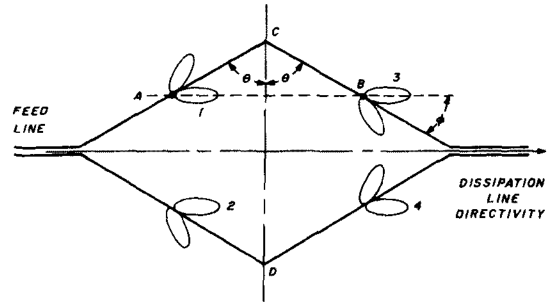

fig. 1. Unidirectional lobing pattern of a rhombic antenna is generated by terminating it in its characteristic impedance through a dissipation line. Lobes 1, 2, 3, and 4 combine their energy when θ, the tilt angle, is equal to 90 degrees minus ϕ.

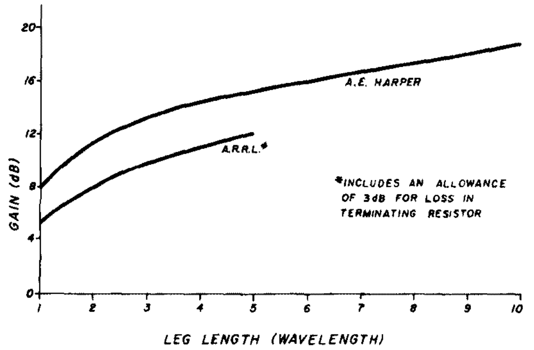

fig. 2. Free space gain of a rhombic antenna over a halfwave dipole for zero wave angle.

tilt angle, θ, shown in the figure, must be adjusted to equal 90 degrees minus the angle ϕ, between the main forward power lobe and the individual leg, which is determined by the antenna length. This ensures that maximum directivity is on a line bisecting the rhombic as indicated.

To obtain the correct phasing of the lobes for maximum radiation in the desired direction, the straight-line distance, AB, between the center of the legs, must be one-half wavelength less than the distance ACB. This follows from the fact that lobe 1 is 180 degrees out of phase with lobe 3. By making the distance between these lobes one-half wavelength (180 degrees) less, lobe 1 will arrive at point B in the correct phase to add to the field of lobe 3, and thus increase the intensity of radiation in the desired direction. A similar action takes place between lobes 2 and 4 on the other side of the rhombic. All other lobes combine to produce a cancellation of radiated energy in the line of the minor axis, CD. Correct termination of the antenna with approximately 800 ohms nonreactive resistance produces an almost infinite front-to-back ratio.

The issue of rhombic gain is a controversial one. **Figure 2** shows gain curves from two sources: *The ARRL Antenna Book,*[2] and *Rhombic Antenna Design,* by A. E. Harper.[3] Both curves are free-space directivity gains of a nonresonant rhombic over that of a dipole and are for *zero vertical angle of radiation.* E. Bruce, the major developer of the rhombic, shows some actual experimental data in his August, 1931, article in the *Proceedings of the Institute of Radio Engineers.*[4] His data shows that in comparison with a halfwave vertical antenna, his three wavelengths on a leg rhombic had a gain of 21 dB 10 percent of the time, to 7 dB 100 percent of the time, and 16 dB 50 percent of the time. Put another way, the rhombic was always 7 dB better than the halfwave vertical, and 10 percent of the time it was 21 dB better. That 21 dB relates to a power ratio of about 130; that is, a 1 kW output transmitter would have an effective radiated power of 130 kW with respect to a dipole — but only 10 percent of the time.

In a detailed article in the January, 1935, *Proceedings of the IRE,*[5] Bruce described experimental data showing that three and one-quarter wavelengths on a leg rhombic had 14 dB gain over a halfwave horizontal dipole at the same height. A Yagi producing 14 dB gain would require 12 elements — a rather large antenna; of course it would be capable of rotation over 360 degrees.

design

For optimum performance, a rhombic antenna should be designed for one frequency or a very small band of frequencies, the pattern for which is best suited to the propagation conditions of the radio circuit. Usually about all that a designer attempts to compute about this system is the characteristics of the main lobe. The enormous labor of computation quickly discourages analysis of a rhombic's complete radiation characteristics. Charts have been provided to assist in suitable designs as shown in the *ARRL Antenna Book,* or Laport's *Radio Antenna Engineering,* figure 3.81.[6] By careful design (and acceptance of less than optimum performance) a rhombic antenna may be made to operate over an almost 3 to 1 frequency range. This means that a fixed rhombic could operate from 3 to 9 MHz, or from 7 to 21 MHz, *or over* any similar frequency range. It will be shown later that a Controlled Vertical Radiation (CVR) rhombic can operate well from its lowest design frequency to as high as practical before beamwidth becomes too narrow for normal use.

To properly analyze a rhombic over a range of fre-

quencies and desired *vertical angle of radiation*, it is necessary to have a method for observing quickly the effects of varying any single parameter in relation to all the others. These parameters are the antenna height above ground, the length of the legs, and the included angle between the legs, called the tilt angle (see **fig. 3**).

The length and tilt angle (**fig. 3A**) are actual dimensions of the antenna proper and affect the free-space pattern. The height of the antenna, however, affects the directional characteristics only through what is called "ground reflection." In **fig. 3B**, part of the radiated power goes directly from the antenna at the vertical angle of radiation, delta. The rest of the radiated power is directed toward the ground at the same angle. If the ground is assumed a perfect reflector, the wave that is directed toward the ground will be reflected, undiminished in strength in the same direction as the original directly radiated wave. If the ground reflected wave arrives at a distant point (your QTH, for example) in phase with the direct (sky) wave, it will reinforce the received signal (voltage). If, however, it arrives exactly 180 degrees out of phase, it will completely cancel it.

In the practical case, neither the maximum of 6 dB reinforcement nor the complete cancellation ever occurs, since the ground is never a perfect reflector. Also, the reflected wave rarely reaches the unreflected wave exactly in phase or exactly out of phase unless the antenna is being used at a frequency exactly that of the design frequency. The effects of the ground reflection will be treated separately from those due to antenna length and tilt angle.

A computer can be used to observe the effect of changing certain parameters. However, it is much easier to see such changes by means of a graphical method Donald Foster described in the October, 1937 issue of the *Proceedings of the IRE*. "If the direction of zero and maxima of K^2 (the radiation function) are plotted on a spherical blackboard with the rhombus at the center, they consist of a coaxial system of small circles, of alternating maxima and zero . . . around one arm of the antenna as an axis, and an identical system of circles around the other arm of the antenna. The angle between the axis of the circles is the angle 2A (see **fig. 3A**) of the rhombus. This pattern on the sphere is ideally suited to representation on the plane by means of the sterographic projection."[7]

While his method probably sounds complicated, it actually summarizes a very simple and easily grasped idea of what happens as the parameters are changed. Antenna enthusiasts will recognize the "zero and maxima of K^2" as the first null and main lobe of the antenna field intensity graph of "Angle with Respect to Wire Axis vs Length of Wire in Wavelength" shown in figure 15 in the second chapter of the *ARRL Antenna Book*.

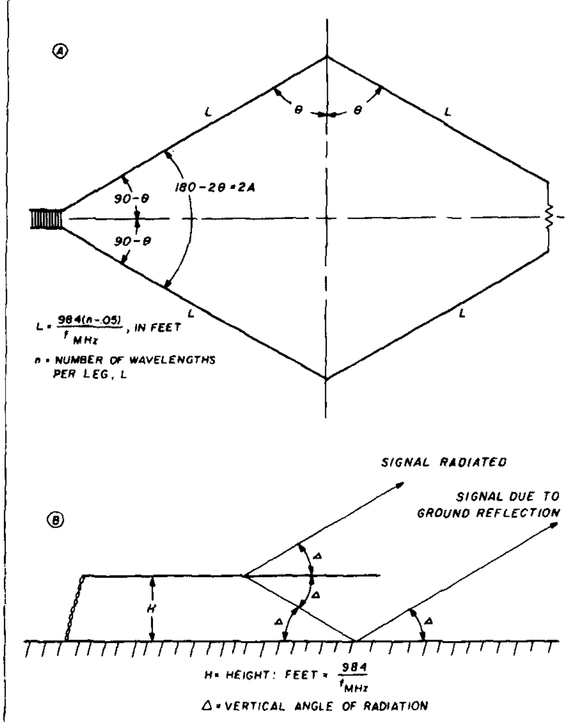

fig. 3. Definition of rhombic antenna parameters. In (A) Length, L, and tilt angle, θ, are actual dimensions of the antenna and affect free-space pattern. In (B), part of the radiated power goes directly from the antenna at angle Δ.

Without going into the mathematics of their construction, the following discussion will help you to make your own stereographic representation.

construction of stereographic overlays

The free-space pattern charts are made by the use of **fig. 4**, which shows the angles of the maxima and nulls in long wires. This is the same as figure 15 of chapter 2 of the *ARRL Antenna Book*, but shows up to the 8th maxima and null instead of just the first.

Step 1. Draw a 6-inch diameter circle and place perpendicular vertical and horizontal lines through its center; refer to **fig. 5**.

Step 2. Lay out 0 to ±90 degree tick marks around the periphery of the circle counterclockwise and clockwise from the right horizontal line intersection of the circle.

Step 3. From the −90 degree position on the perimeter of the circle, draw straight lines to 0, ..., 80, 90 degrees and label the points where they intersect the horizontal line as 0, 10, ..., 70, 80 degrees. For ease of viewing **fig. 5**, only the 20, 50, and 70 degree lines are shown.

The labeled points represent the vertical angle of

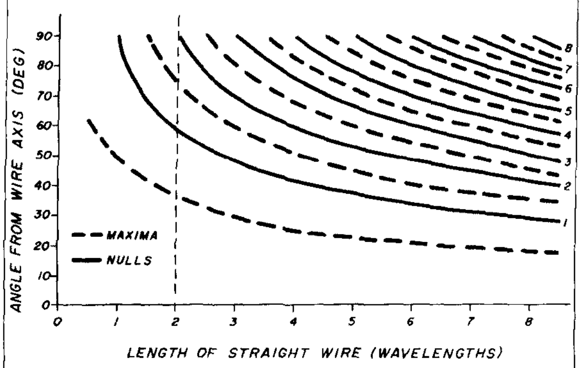

fig. 4. Angles of maxima and nulls in long wires carrying standing waves.

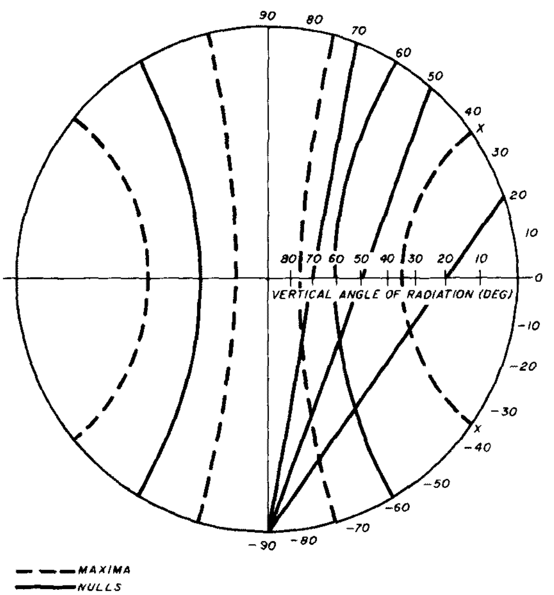

fig. 5. Stereographic map of the radiation pattern for a two-wavelength straight wire in free space.

radiation of the rhombic. It will be necessary for future work to have this as an overlay for a scale for determining the vertical angle of fire of other rhombic designs.

Step 4. Determine the number of wavelengths to be analyzed (2 wavelengths, for example).

Step 5. Using **fig. 4**, draw a vertical line up from the 2 wavelength point to the curve. Read off the indicated value.

Step 6. Tabulate the angle of maxima and nulls from Step 5; maxima at 36 and 75 degrees, nulls at 60 and 90 degrees.

Step 7. Place a mark on the circle perimeter at ±36 degrees. Draw a dotted circle through those two points and the 36 degree point on the horizontal line. The center of that radius must lie on the horizontal line extended to the right of the circle. That dotted line represents the first maximum of a radiating wire two wavelengths long. Repeat for ±75 degrees, which represents the second maximum. Repeat the same two angles from the left side of the chart, which represent the reverse direction of fire.

Step 8. Place marks at ±60 and 90 degrees and draw solid-line curves, which represent the first and second nulls respectively. Repeat for the reverse direction.

You now have a one-leg pattern of a two-wavelength long rhombic. You will need two of them for analysis, as will be explained.

A ground reflection overlay is also needed and is made as follows.

Step 1. Draw a 6-inch diameter circle with a perpendicular horizontal and vertical line through its center; see **fig. 6**.

Step 2. Determine the number of wavelengths above ground the antenna is to be placed (one wavelength, for example).

Step 3. Using **fig. 7**,[8] tabulate the null and maximum vertical angles of radiation for the chosen height. For one wavelength (360 degrees) we have 15 degrees, 48 degrees maximum and a 30 degree null.

Step 4. Place the scale for determining the vertical angle of fire of the rhombic under the 6-inch circle and place a mark on the horizontal line at the 15 degree, 48 degree, and 30 degree vertical angle of fire points.

Step 5. Draw dotted-line circles through the 15 and 48 degree marks using the center of the 6-inch circle so as to produce concentric circles. Draw a solid-line circle through the 30-degree line in the same manner. The dotted circles represent the first and second maxima, and the solid line represents the first null.

The three stereographic maps are all that are required to design a two wavelength on a leg rhombic mounted one wavelength above the ground. All other maps are made the same way for different leg lengths and heights.

During World War II Richard Bluhm, W2KXD, adapted Foster's graphical method to make it practical for use by the average person. In an unpublished paper[9] written in 1944, he provided a means of rapidly designing horizontal rhombic antennas using Foster's stereographic overlays. Even though the data obtained by his method is not precise, results obtained during wartime erection of rhombics by the military bear out mathematical calculations with excellent accuracy.

The design of my rhombic is based on W2KXD's method. In discussions with him, we both felt that his

stereographic charts (almost 50 in number) should be available to Amateurs interested in designing and constructing rhombics.*

Figure 5 is the stereographic representation of the free-space radiation function of one leg of a rhombic antenna. The length of this leg is two wavelengths. Let's review it for emphasis. Looking from right to left on **fig. 5**, there is first a dotted line, then a solid line, then a dotted line and so forth. The dotted curves represent the maxima circles described by Foster. The solid curves represent the zero circles. A drawing identical to **fig. 5** is then superimposed on the drawing shown in **fig. 5**. Each represents the radiation function of a two-wavelength leg of a rhombic antenna.

Suppose it is desired to have a tilt angle of 70 degrees. From **fig. 3A** we can calculate 2A = 40 degrees. By rotating the superimposed drawings so that a 40 degree angle is realized between the axes of the two legs, we obtain the actual free-space radiation pattern of a rhombic antenna with two wavelength legs, and a tilt angle of 70 degrees; see **fig. 8**.

By studying **fig. 8**, it can be seen that the first dotted lines of the two drawings intersect at point X; that is the main lobe. Next, consider the first dotted line of the lower leg and note that it intersects the second dotted line of the upper leg at point Y. The second dotted line of the lower leg intersects the first dotted line of the upper leg at point Z. Other points of intersection are at points A, B, C, D, E, F, and G as shown in **fig. 8**. These intersection points of the dotted circles represent points of maximum radiation, or lobes, of the antenna.

A line drawn from the center of the figure through point X is extended to the edge of the great circle. This line is now called "the axis of the antenna," and is the line in which the strongest lobe of the rhombic lies. The strongest, or main lobe of a rhombic will always fall exactly midway between the two legs of the antenna if it is designed correctly.

The next step is to number the dotted curves at the periphery of the circle for ease of handling. The dotted lines of the upper leg are numbered 1, 2, 3, and 4, starting with the lower end of the first dotted line and going clockwise. The dotted lines of the lower leg are also numbered 1, 2, 3, and 4, but starting at the upper end of the first dotted line, and going counterclockwise.

When a Number 1 curve intersects another Number 1 curve, the resulting point is that of maximum radiation of the antenna. Other intersection points, called minor lobes, do not reach the level of the (1,1) intersection point. For instance, (refer to **table 1**), a Number 1 dotted curve intersecting a Number 2 dotted curve, points Y and Z on **fig. 8**, gives a lobe which is 10.6 dB lower in level than a (1,1) intersection. A (2,2) intersection point A on **fig. 8** is 21.1 dB lower

*A complete set of approximately 50 8½ by 11 inch overlays may be obtained from the author. These overlays are in photocopied form and will have to be made into transparencies for actual use. — **Editor.**

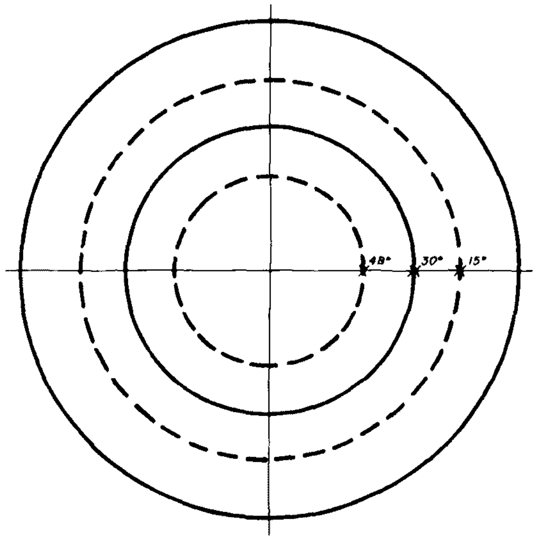

fig. 6. Ground interference pattern; antenna height one wavelength.

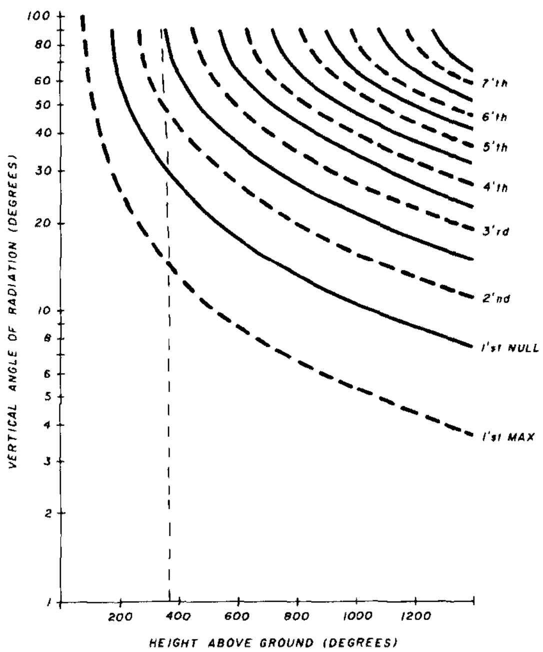

fig. 7. Vertical radiation angle vs. maximum and null angles for various antenna heights in electrical degrees.

table 1. Decibel differences between main lobe and subsequent minor lobes on rhombic antennas.

	rhombic radiation lobes										
	1	2	3	4	5	6	7	8	9	10	11
1	0										
2	10.6	21.1									
3	15.1	25.6	30.10								
4	18.0	28.6	33.10	36.0							
5	20.2	30.8	35.25	38.2	40.40						
6	21.9	32.5	37.00	40.0	42.10	43.9					
7	23.4	34.0	38.50	41.1	43.60	45.3	46.8				
8	24.6	35.2	39.70	42.6	44.80	46.6	48.0	49.3			
9	25.7	36.3	40.80	43.7	45.90	47.7	49.1	50.4	51.4		
10	26.7	37.3	41.75	44.7	46.90	48.6	50.1	51.3	52.4	53.40	
11	27.6	39.2	42.60	45.6	47.75	49.5	51.0	52.2	53.3	54.25	55.1
	power difference (dB)										

in level than the (1,1) intersection. The (4,4) intersection, point G is 36 dB lower, and so forth.

As the number of wavelengths on a leg increases, the number of dotted curves increases. For a composite overlay of two ten-wavelength legs, there will be twenty dotted curves per leg. **Table 1** gives levels only up to the eleventh curve since the minor lobes beyond this point are so weak in comparison to the main lobe as to be negligible. If at any time a solid curve intersects two dotted curves at or near their intersection, the lobe made by these curves intersecting will be cancelled or considerably reduced. That holds true in all cases. For instance, the (1,2) intersections, or the Y and Z points in **fig. 8**, are very close to the outside circle, which represents the horizon. Since these two lobes are depressed to just above the horizon, they may be considered as absorbed by surrounding hills or buildings, so that they will be of little use.

To find the vertical angle of fire of each of the lobes, **fig. 9** is superimposed on the drawings. **Figure 9** is the scale used for determining the vertical angle of fire for the various lobes of the rhombic antenna. When **fig. 9** is placed coincident with **fig. 8** (all figures are transparent), the main lobe, point X, will be at a 30 degree vertical angle of fire. Points Y and Z have about 5 degree vertical angle of fire, which is too low to be usable except at the higher frequencies. Point A has about a 73 degree angle, points B and C about 90 degrees, and so forth.

The great thing about this stereographic method of analysis is that angle 2A between the two legs can be easily changed. As the angle is made greater, the two legs move apart, the main intersection travels out toward the horizon, and the vertical angle of radiation becomes less. Angle 2A can be increased until the intersection of the number one dotted curve reaches the horizon with a vertical radiation angle of 0 degrees. Further separation between the two legs causes the main lobe to split into two lobes. Although not obvious, the two split lobes are excessively sharp horizontally; yet despite this sharpness, the gain must necessarily be very low. Some of the minor lobes will have magnitudes as great or perhaps greater than the main lobe. The energy of the system is leaking out through other lobes in other directions rather than being concentrated in the main lobe. It might be difficult to discover this situation by arithmetical computations, but it is quicky observed by using stereographic charts. For a fixed rhombic beam, the upper frequency use is limited at the point at which the beam splits.

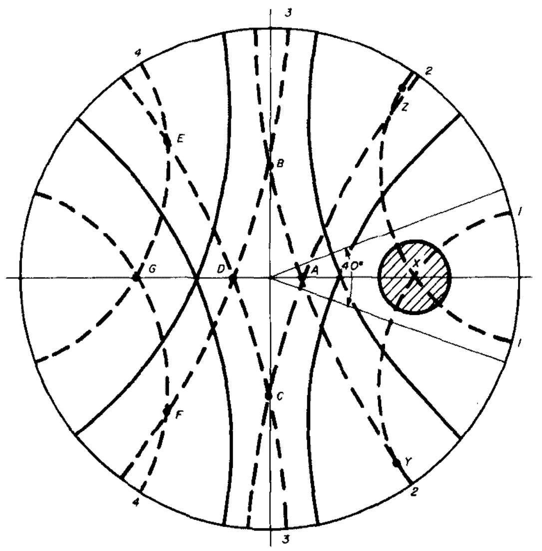

fig. 8. Free space pattern, two-wavelength. Note: the horizontal and vertical lines are for reference to the antenna plane only and are not seen in the overlays.

ground reflection effects

We must now consider the ground reflection or

ground interference effects on the free-space pattern of the two wavelength rhombic discussed above. To keep it simple, an antenna height of one-half wavelength will be used; it has only one reflection agent. **Figure 10** shows the ground pattern, which is the dotted circle, for one-half wavelength high antenna superimposed on **fig. 8**. Dotted lines represent an in-

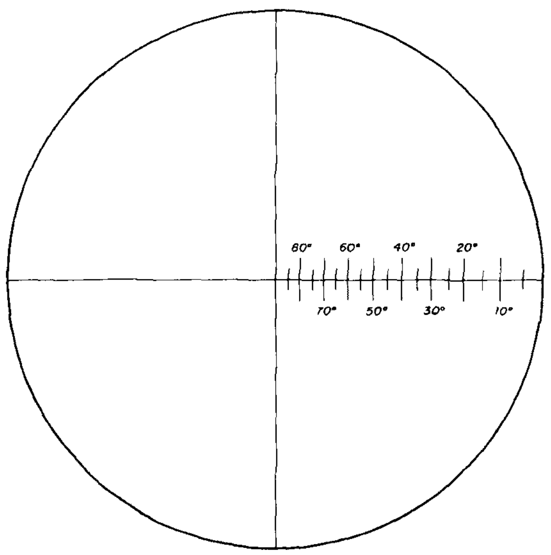

fig. 9. Scale to determine vertical angles of fire on rhombic antennas.

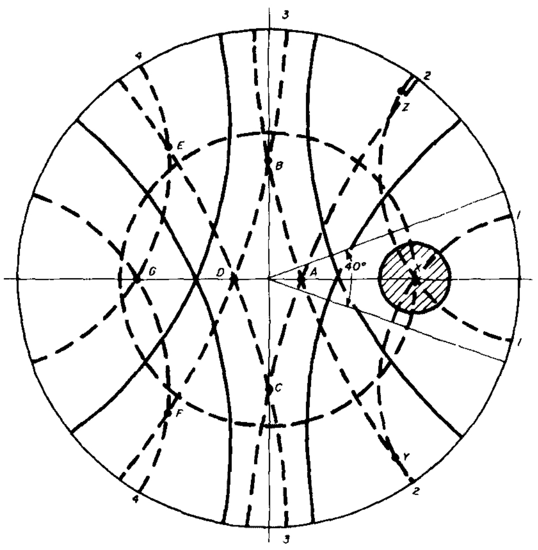

fig. 10. Superimposition of two-wavelength rhombic free-space pattern and one-half wavelength height ground interference pattern.

phase reflection reaching the antenna, whereas solid lines represent cancellation. The one-half wavelength height has no out-of-phase reflection radiation.

Note that the dotted circle intersects the free space antenna pattern *exactly at point* X, which is the main lobe point. (This happened only because the problem was done before writing this article, of course). That means the in-phase reflection has arrived to reinforce the main lobe. As noted previously, an additional 6-dB reinforcement of the main lobe has occurred. Points Y and Z of **fig. 10** are quite distant from the dotted circle and therefore are not reinforced. That means the main lobe has now increased to 16.6 dB instead of only 10.6 dB stronger than the next two strongest lobes (Y and Z) *merely by choosing the correct antenna* height.

Point A, the next strongest lobe, has not been reinforced by ground reflection and consequently has also been reduced with reference to the main lobe. Points B, C, E, F, and G are relatively close, however, to the reflection circle. While they will not be reinforced by 6 dB, since they are not exactly on the reflection circle, they have been reinforced by perhaps 4 dB so they have only decreased by a matter of, say, 2 dB with respect to the main lobe.

The wavelength height of the antenna may be raised by using higher towers for a given frequency or by increasing to a higher frequency with fixed antenna height. As the wavelength height of the antenna is raised, the number of reflection circles increases, and a number of solid lines representing ground interference appears on the stereographic overlays. For example, at a height of one wavelength, two reflection circles separated by an interference solid-line circle appear. At two wavelengths, there are four reflecting dotted circles and three interference circles. Different ground interference pattern overlays are therefore required.

design of 20-meter rhombic

Now we have an idea of what the use of the overlays can do for us in designing rhombics. Let's apply that information to the design of a rhombic for use on the 20-meter band. We want it to be four wavelengths on a leg and have it one wavelength high. The problem is to determine the best tilt angle for these conditions. Note that this is not the proper way to start a rhombic antenna design. The proper way is to first determine the radio circuit path desired, and therefore the desired vertical angle of radiation between the transmitting and receiving stations; review reference **8**. Then design the rhombic to include that radiation angle. However, the stated problem will best review the use of the stereographic overlays, permitting you to do what you really want to do.

Superimpose two "leg pattern — four wavelengths"

and one "ground interference pattern — one wavelength." Rotate the two leg pattern overlays so that their (1,1) curves intersect at the outermost dotted circle of the ground interference chart. Measuring the angle between the two legs, we find it to be 42-1/2 degrees, which is the angle 2A of **fig. 3A**. The tilt angle, θ, will be calculated to be 68-3/4 degrees. If the scale for determining the vertical angle of fire, **fig. 9**, is placed on the other charts, it shows the vertical angle of radiation for our rhombic to be 14 degrees. **Figure 11** shows the resulting composite of the overlays, first maxima only.

All the essential design factors of the rhombic have been found. They are: leg length = four wavelengths; height = one wavelength, tilt angle = 68-3/4 degrees, angle of fire = 14 degrees. The relative strength of the minor lobes and the front-to-back ratio can be found by referring to **table 1**. The front-to-back ratio in this case is about 51 dB. This is found by determining the intersection of the last two dotted curves which lie on the line extending in back of the main lobe. These will always be the dotted curves nearest the left-hand edge of the solid circle of the leg patterns (not shown in **fig. 11**). In this case they are the Number 8 curves counting from right to left on the leg patterns. Referring to **table 1**, the lobes produced by the intersection of the Number 8 curves or (8,8) point is 49.3 dB weaker than the main or (1,1) lobe. The main lobe has a ground reinforcement of 6 dB. Because the ground reflection pattern does not pass directly through the (8,8) lobe, but only near it, a ground reflection of about 4 dB will accrue. The difference between these two reinforcements is thus 2 dB, which further increases the front-to-back radio to 51.3 dB. In round numbers, this is about 51 dB. The vertical angle of radiation of the (8,8) lobe is about 22 degrees.

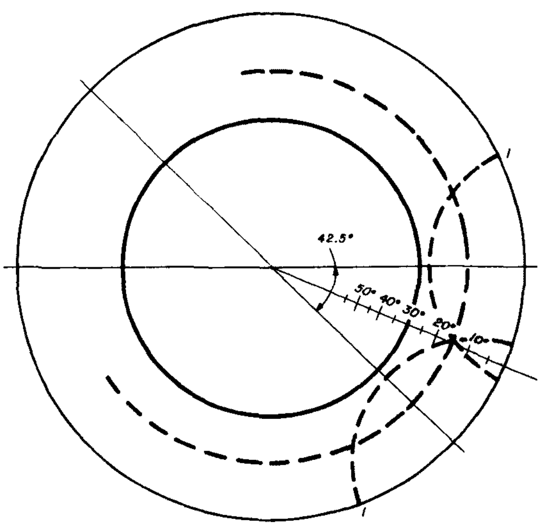

fig. 11. Overlay of two two-wavelength leg patterns, a ground interference pattern one wavelength high, and scale for determining vertical angle of radiation. Intersection with the first maximum ground interference line of first maximum (1,1) produces a 2A = 42-1/2 degree angle, and a vertical angle of radiation of 14 degrees.

Note that there is a (7,7) rear lobe falling exactly on the second ground reflection circle. However, the vertical angle of radiation is about 48 degrees, which will probably be lost into space at 14 MHz.

determination of vertical and horizontal radiation patterns

Figure 12 duplicates **fig. 8**, but with many lines eliminated to make the figure less "busy." Radial lines have been drawn from the center of the figure through each lobe point and extended a distance beyond the horizon circle. Starting at the axis of the antenna, the angle between the lobes is measured and recorded. These radial lines at the recorded angles are then reproduced on polar coordinate paper; see ***fig. 13***. The radial lines at their correct angular displacement from the axis of the antenna are marked with their corresponding dotted line intersections. That is, the axis of the antenna will be marked with (1,1) and (2,2) because the Number 1 dotted curves and the Number 2 dotted curves have their intersections on that line. The other lines are similarly marked.

Next the ground interference pattern for an antenna height of one-half wavelength as shown in **fig. 10** is examined and the strength of all lobes as previously discussed (**table 1**), are tabulated. They are:

(1,1) =	0 dB	(3,3) =	−30.1 dB
(2,1) =	−16.6 dB	(4,3) =	−35.1 dB
(2,2) =	−21.1 dB	(4,4) =	−38.0 dB
(3,2) =	−27.6 dB		

The scale on the polar coordinate paper is then laid out from 0 dB through 50 dB in 10-dB increments; see **fig. 13**, and the lobes are plotted with reference to these circles.

This is only a relative pattern because it does not take into account the effect on the pattern of the different vertical angles of fire. For instance, the (2,1) lobes are greatly attenuated because of their low angle of fire, resulting in absorption by surrounding hills or buildings if in the vicinity of the antenna. However, the pattern does show the relative strengths of the peaks of the lobes in the horizontal plane and should prove very useful.

The determination of the strengths of the lobes at any point on, with the exception of the peak, is not readily determined through the use of the stereographic projection. The horizontal angle covered by the major lobe may be roughly found by drawing a

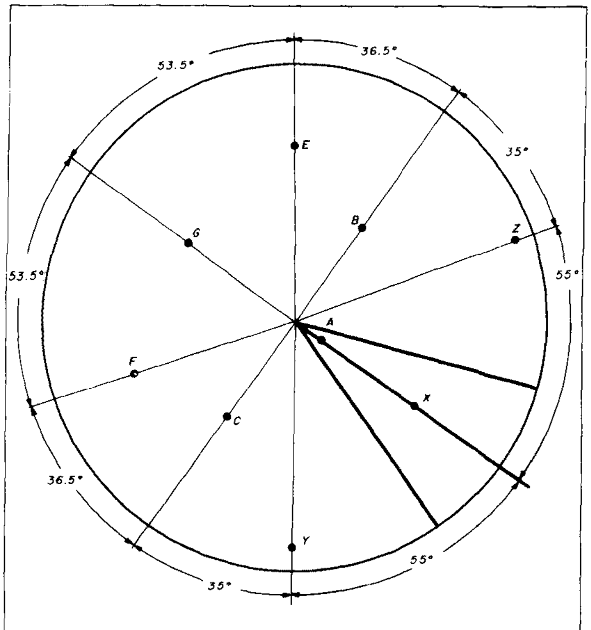

fig. 12. Same as fig. 8, but with intersection points only shown with angular displacement of intersections from main lobe direction.

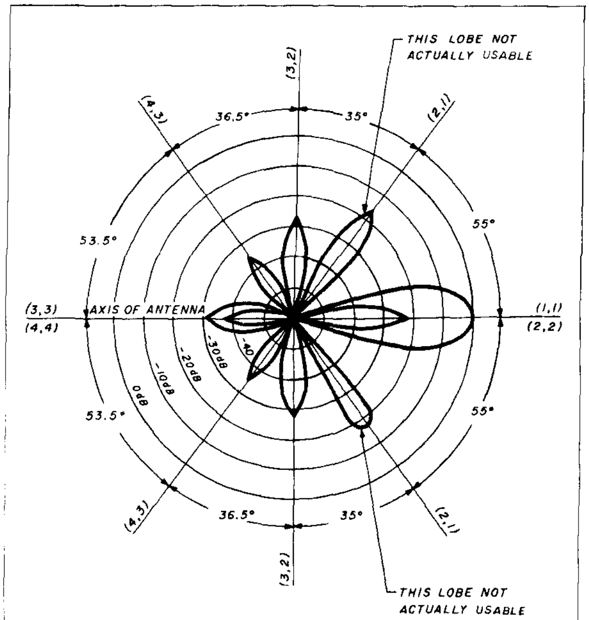

fig. 13. Horizontal plane pattern for two-wavelength rhombic antenna. ϕ = 70; antenna height = one-half wavelength.

circle with the center at the (1,1) intersection, (or at the center of the spherical triangle formed by the intersection of Number 1 curve and the ground interference circle, depending on whether all three curves intersect in one point or not) and a radius equal to one-half the distance between the intersection (or spherical triangle) and the nearest solid null curve. Lines tangent to this small circle drawn from the center of the large circle will form an angle which is a rough estimate of the usable horizontal beam width. On **fig. 8**, the small shaded circle is the circle mentioned above. The beamwidths of the minor lobes are not readily obtainable, but this should not prove objectionable since the major lobe is the only one which is used in most of the cases.

The vertical plane diagram is constructed in a similar manner as the horizontal pattern, except that **fig. 9** is used to determine the vertical angles. These are:

(1,1) = 29 degrees	(3,3) = 75 degrees
(2,1) = 7 degrees	(4,3) = 27 degrees
(2,2) = 75 degrees	(4,4) = 30 degrees
(3,2) = 35 degrees	

This gives some interesting results. Above about 7 MHz the (2,2) and (3,3) lobes may be considered useless since they will penetrate the ionosphere at such high angles. As previously stated, the (2,1) lobes are radiated at such a low angle as to be useless on all but extremely local signals or at extremely high frequencies.

Returning to the vertical plane pattern, radial lines are again laid out, only this time the vertical angles of fire are used in place of the horizontal radiation angles. All those lobes falling to the left of the center of the circle are plotted to the left and all those falling to the right are plotted to the right. Those falling to the left will be (3,3), (4,3), (3,2) and (4,4). Those falling to the right will be (1,1), (2,1) (3,2) and (2,2). Since the (3,2) lobes fall exactly on the center line of the circle, see **fig. 8**, (axis of the antenna being horizontal) one will be plotted to the left, and one to the right. This is not a strictly accurate geometrical layout of the pattern but will suffice since the determination of the relative strengths of the lobes in the vertical plane is all that is desired. The levels determined for the horizontal plane pattern may be used, without change, for the vertical plane pattern. The complete vertical plane pattern is shown in **fig. 14**.

Through the use of different height curves, leg length curves, and tilt angles, unwanted lobes may be eliminated or effectively reduced and desired lobes may be reinforced. The use of leg lengths longer than eight or ten wavelengths is inadvisable because of the subsequent reduction in width and height of the radiated lobes. A reduction of this sort is conducive to fading and makes the aiming of the antenna extremely critical.

A compass rose may be superimposed on the drawing as an aid in determining the angle between the legs of the antenna and the angles between the main lobe and the minor lobes. The use of an angle between the

table 2. Beamwidth as a function of leg length. Beamwidth is the point where power is 3 dB down from the maximum power point.

leg wavelength	beamwidth (degrees)	number DXCC countries within beamwidth (from North Carolina): forward direction	reverse direction
1	30	84	4
2	25	77	4
4	17	63	4
6	10	43	4
8	8	40	4
10	6	35	3

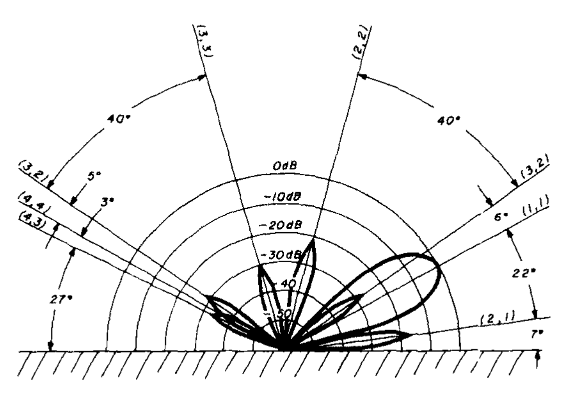

fig. 14. Vertical plane pattern for two-wavelength rhombic antenna. ϕ = 70; antenna height = one-half wavelength.

legs, such that the two Number 1 curves do not intersect, should be avoided, because this will effectively eliminate the major lobes, which is the most effective source of power from the rhombic antenna.

multiband operation

One of the most useful features of a rhombic antenna, as previously mentioned, is its ability to operate efficiently over a wide frequency range. The two-wavelength rhombic illustrated in **fig. 8** will be used as an example. Suppose it is desired to operate this antenna on 4 MHz. Its legs will then be two wavelengths long and its height will be one-half wavelength. On 4 MHz, using the formulas of **fig. 3**, two wavelength legs will be 485 feet long, and the height of the antenna will be 123 feet. Now, if the operating frequency is made 8 MHz, the length of the legs will be four wavelengths long at this frequency, and the antenna height will be one wavelength. Suitable overlays for the leg length and antenna height for this frequency are now set up, retaining the 70 degree tilt angle previously used. The major lobe will now be found to have an angle of fire equal to 15 degrees which is approximately the optimum angle of fire for 8 MHz. The horizontal azimuthal angle covered by the major lobe has now been reduced to 17 degrees instead of the original 25 degrees. The antenna will work practically as well, therefore, on 8 MHz, as it does on 4 MHz. The only change worth noting is the reduction of the beamwidth, since the 15 degree angle of fire is about optimum for 8 MHz, as is the 30 degree angle for 4 MHz.

Suppose the frequency were now increased to 12 MHz. The legs will now be six wavelengths long and the height will be one and one-half wavelengths. The pattern is again set using the appropriate overlays. It will be noted that the two (1,1) curves, depressed to the horizon, form a spherical triangle with the first ground interference circle. While this materially reduces the strength of the major lobe, it is still usable. The angle of fire is taken from the center of the spherical triangle and is found to be about 7 degrees. While this is rather low for 12 MHz, it will work fairly well in locations where the antenna is well out in the clear and away from any trees or buildings. This rhombic antenna, therefore, may be said to be extremely effective over a 2 to 1 frequency range (4 - 8 MHz) and fairly effective over a 3 to 1 frequency range (4 - 12 MHz). It would work over any 3 to 1 frequency range, 6 to 18 MHz, 8 to 24 MHz, or any similar range. Of course appropriate leg lengths and heights would have to be used for two wavelength legs and one-half wavelength for the lowest frequency.

beamwidth

The beamwidth of a rhombic is generally a function of the leg length. **Table 2** indicates beamwidth versus leg length. Beamwidth is defined as the angle where a 3 dB loss has occurred from the maximum power point. From Salisbury, North Carolina, with the rhombic pointed at approximately 47 degrees from north, which is the bearing for London, England, many countries can be worked as shown in **table 2**.

By switching the feedpoint to the rhombic including the terminating resistor, to the opposite end of the

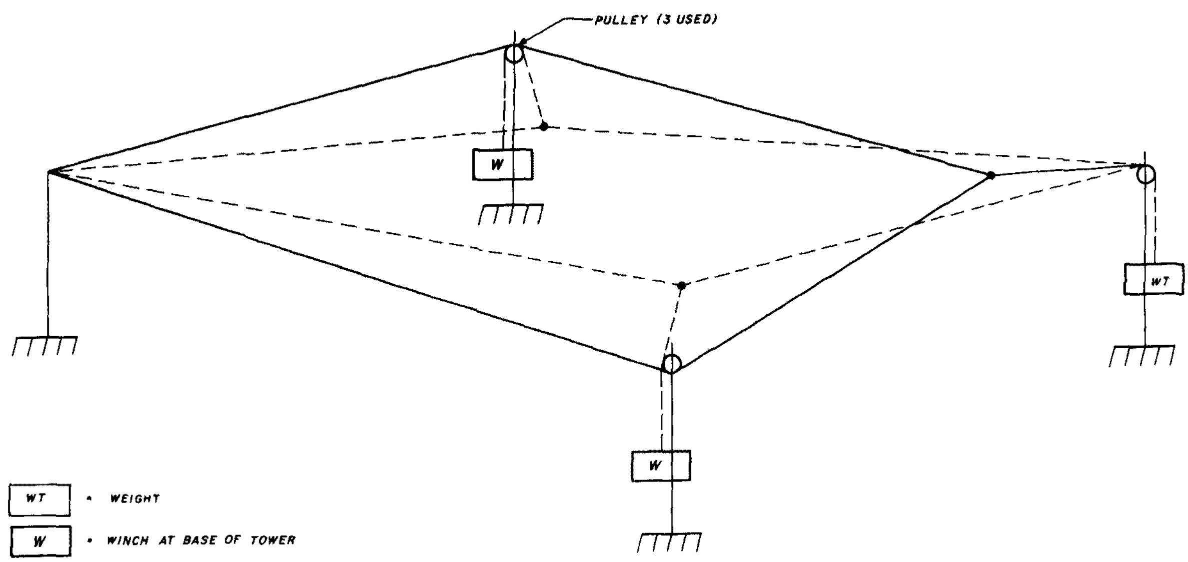

fig. 15. A controlled vertical radiation (CVR) rhombic antenna showing extremes between lowest frequency use (solid lines) and highest frequency use (dotted lines.).

rhombic, its direction of fire will be reversed 180 degrees. **Table 2** also shows the number of DXCC countries workable within the stated beamwidths in the reverse direction — not very many. It is, however, a "pipeline" through Mexico, Pitcairn Island, Clipperton, and MacQuarie. With a rhombic on Japan, South America should be blanketed in the reverse direction. Unless you are interested in a particular point-to-point radio path, rhombics longer than 10 wavelenths are too narrow for general use.

controlled vertical radiation rhombic

Now that you know how to design a rhombic antenna, let's move on to a more specific aspect of design, the controlled vertical radiation (CVR) rhombic. Bruce and Beck, in the April, 1935 *Proceedings of the IRE*[10] described experiments made with a steerable rhombic during reception of transoceanic shortwave signals. The first and last I'd read about it in Amateur Radio was an article in April, 1937, issue of *QST*.[11] In this account, W6AUX and W7CNX reported on their operation of a CVR rhombic in the 20-meter band. While not claiming anything new, I am expanding the CVR principle to provide a rhombic that can operate at a design efficiency anyplace in the Amateur bands with a limitation only on the minimum acceptable beamwidth.

A CVR rhombic is simply a rhombic whose shape may be changed by physical means; see **fig. 15**. By having pulleys on the side towers and one end tower,

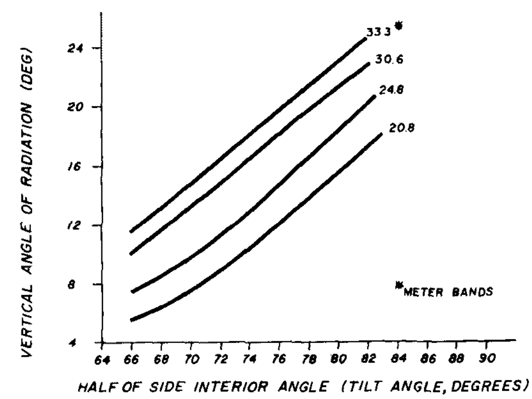

fig. 16. Change of vertical angle of radiation with change of tilt angle for various operating frequencies. (By E. Bruce in 1935.)

the tilt angle of the rhombic may be changed and set to any desired number of degrees.

Let's see what Bruce and Beck say about that. They were studying rapid fading in radio circuits, and the possible cause being the interaction of different components of a radio signal having different transmission times. Their past observations had indicated that fading was affected by the directivity of the receiving antenna. *Tests in 1934 had shown that a* greater degree of angular spread between multiple path waves exist in the incident vertical plane than in

the horizontal plane. So they devised a rhombic of the type of **fig. 15** and ran extensive tests. Their article concludes, "It is believed that the results, discussed in this paper, demonstrate that sharp angular discrimination is a basically sound method of combating selective fading."

Of greater interest to Amateurs is not the minimizing of fading, but the fact Bruce's and Beck's tests showed that the vertical *angle* of radiation from a rhombic can be varied 12 to 14 degrees for a given frequency. While it's possible to do that with a Yagi antenna by raising and lowering its tower, how many hams would want to do that? The compromise — a good one — is to have a high Yagi for long-haul or band-opening contacts, and a low one for staying within the skip zone into Europe when the band is wide open. However, we're talking about a rhombic with superior gain over a Yagi when the rhombic is operating at its peak.

Figure 16 is a copy of the Bruce/Beck curve showing steerability, at several wavelengths, of the horizontal rhombic antenna used for fading reduction studies. We can call it the vertical radiation angle versus the rhombic tilt angle. Think about setting your rhombic on 20 meters for a 7 degree vertical angle of radiation as sunrise approaches to get real long haul or early band openings, and then as the day continues, changing the radiation angle to 12 degrees or more to put a commanding signal into Europe when the band is fully opened. When motorized, it would be possible to tune the antenna for maximum received signal strength from the desired location.

In the 1937 *QST* article, using the same idea, Moore and Johnson concluded the following:

1. *That there is an optimum angle in the vertical plane for transmission as well as reception.*

2. *That the optimum angle for transmission and reception are close together although not necessarily coincident.*

3. *That there is, under normal conditions, only a very limited region in the vertical plane in which useful radiation takes place, and that energy directed into any other region in the vertical plane is largely wasted.*

4. *That the optimum angle of transmission changes from time to time with changes of seasons and conditions, but that there is no material change within a short interval of time.*

5. *That controlled directivity in the vertical plane is relatively more important than directivity in the horizontal plane.*[11]

Now that rotatable arrays are the accepted thing, the fifth claim is debatable. However, in the 1930's one would have assumed Bruce and Beck meant that the proper vertical angle of radiation to a given point is more important than the gain of the antenna; gain and directivity, at that time, seemed to have been synonymous. A very high gain antenna whose vertical angle of radiation over-shot the desired reception point would be a poor performer in comparison to a dipole whose vertical angle of radiation was such as to give maximum reception at the receiving point.

From earlier discussions, we have learned that the vertical angle of radiation does change as we vary the tilt angle of the rhombic. That change is very easy to see with the stereographic overlays. Unfortunately for the earlier investigators, Foster did not publish his works until October, 1937.[7]

We must not be left with the impression that we are getting something for nothing when we change the vertical angle of radiation by tuning the tilt angle. If you recall the analysis section above, you will remember that the tilt angle during design is adjusted to fall on a dotted circle of the ground interference pattern to give a 6 dB boost from the ground reflected signal. By tuning the tilt angle during operating periods, that ground reinforcement deteriorates. However, this is where point five of the Moore/Johnson[11] conclusions becomes important; controlled directivity in the vertical plane is more important *than gain in* the horizontal plane. Since we can tune the tilt angle for maximum received signal, the law of reciprocity of transmitted/received signals says we are at the best operating conditions for the radio path in use.

The most interesting thing about being able to change the configuration of the rhombic is the ability to tune the antenna to the operating Amateur band desired. It was earlier stated that a fixed rhombic can be made to work reasonably well over a range of frequencies of 3 to 1. As the frequency gets higher, the vertical angle of radiation gets lower until at some frequency the main lobe splits and the rhombic no longer has high operating performance.

The CVR rhombic can be adjusted for peak performance at any Amateur band. For example, if for a given arrangement, the antenna frequency is increased to the lobe splitting point, it is necessary only to lengthen the overall configuration to raise the vertical angle of radiation and bring the split lobes together again at the higher frequencies.

RF feed to a rhombic

This discussion of feeding RF to the rhombic is based on the understanding that the antenna will be terminated in its characteristic impedance. By so terminating it, we can take advantage of the excellent front-to-back ratio that distinguishes this antenna from other types, as discussed earlier. The method of termination will be discussed later.

The antenna input impedance changes with frequency even when terminated. Various authorities

show that an impedance change occurs from as much as 850 ohms to 600 ohms over a frequency range of 4 to 23 MHz. However, because of the relatively small percentage change, the worst SWR based on a center impedance of 750 ohms would be 1.25:1. So the problem boils down to getting from the transmitter output of 50 ohms to the antenna's 750 ohms (see **fig. 17**).

It can be seen that the main transmission feeder line is a 600 ohm, two-wire open line, with provisions to feed either end of the rhombic antenna. A switching arrangement at the center of the antenna permits exchanging the RF feedline and dissipation line to allow remote switching of direction of fire while maintaining a high front-to-back ratio in the chosen direction.

You can get to 600 ohms from 50 ohms immediately by using a 12:1 ratio balun. Barker and Williamson makes a 5 kW 12:1 balun; you can also wind your own. Six hundred ohms for the main transmission line was advisable in my case because of the availability of a 118 watt, 600-ohm type CX. The Globar Division of The Carborundum Company makes a non-inductive resistor that can be used in conjunction with the dissipation line as the termination resistance.

Impedance changes from 750 ohms to 600 ohms are required to get to the 600-ohm line from the two ends of the rhombic. A transmission line whose characteristic impedance is gradually tapered from one value to another may be used as a coupling transformer providing the change in impedance along the line is sufficiently gradual.

When a tapered-line transformer with a minimum length is desired, the characteristic impedance must be tapered exponentially between the two limiting values. One can avoid complicated design computations by using an exponentially tapered line section at least one-half wavelength long at the lowest frequency to be transmitted and connecting it directly between the antenna and the transmission line. Such a line was used between the 750-ohm antenna input and the 600-ohm main transmission line. Since I wanted to use the rhombic on 80 meters, a half-wave exponential line of 137 feet in length was constructed for each end of the rhombic.

rhombic termination

If you don't wish to reverse the direction of radiation of the rhombic, a non-inductive resistor may be installed directly at the far end of the rhombic. The power rating of the resistor should be at least one-third of the power going into the antenna; two-thirds of the input power is radiated before it reaches the far end. For example, if the power to the antenna is 500 watts, key down, the terminating resistor should be able to dissipate about 170 watts.

An alternative is to use a balanced lossy line of high dissipation rating. I used a 600-ohm dissipation line

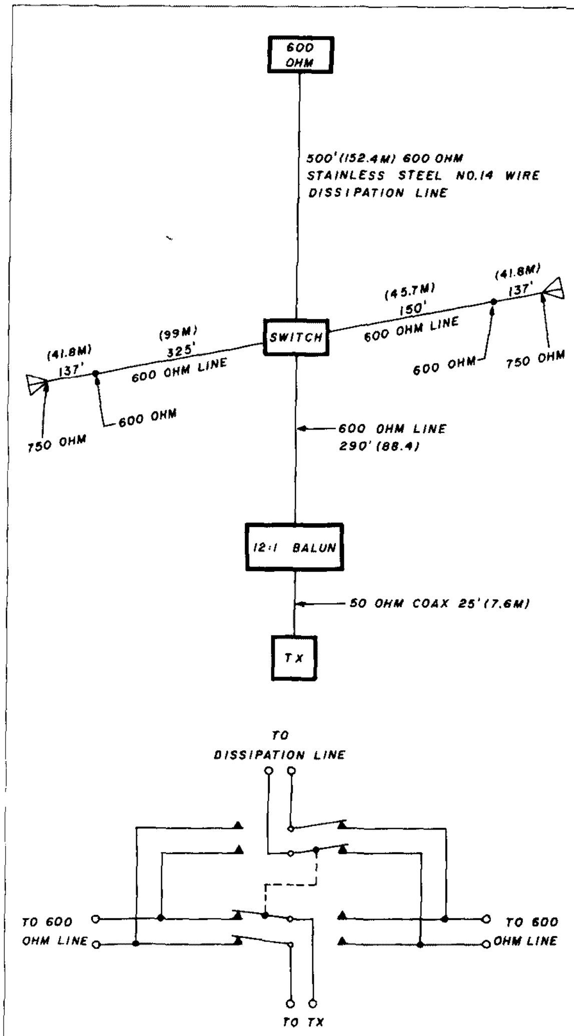

fig. 17. (A) Matching to the rhombic and changing its firing direction is accomplished by using a 12:1 balun, transmission line switch and dissipation line. (B) Transmission line changeover requires an effective 4 PDT switch.

(No. 14 Stainless Steel wire) 500 feet long. The 600-ohm, 118-watt non-inductive resistor terminates the line. As seen in **fig. 17**, remote controlled switching circuits permit swapping of the transmitting and dissipation line to permit reversing the direction of fire of the rhombic.

motorized configuration changer

Figure 15, previously referred to, is a very simplified picture of how to change the configuration of the rhombic. However, because the motorizing of the configuration change by remote means was the most difficult part of the project to develop successfully, some guidelines may be useful.

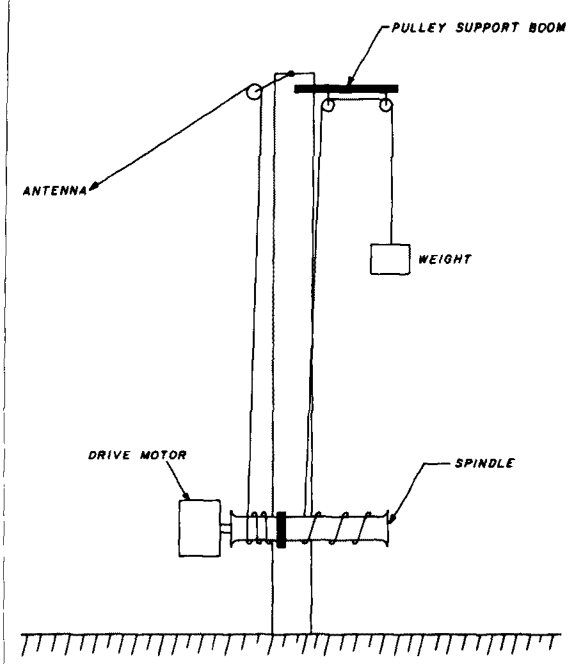

fig. 18. Counterbalancing system to reduce motor drive load.

The first attempt to motorize the configuration was done by a motor at the far end of the array with the weights on the side tower to maintain proper tension in the antenna as the configuration was changed. Don't do it that way! As the rhombic becomes "longer," its legs approach a straight line. The force required to pull the rhombic becomes increasingly greater and a very large motor is required. A better way is to pull the legs in from the side as depicted in **fig. 15**. An analogy is the difficulty in tightening a violin string as opposed to the ease of plucking it from the middle.

Two identical motors — of only modest power — are required. To further reduce the motor power requirements, a counter-balancing arrangement should be used; see **fig. 18**. Using such a system permits a lower motor output because it must only overcome the difference in antenna force versus the counterweight force as the system moves off its equilibrium point, a point of zero motor output requirement.

The drive motors have to be capable of reversing and require a gear reduction to not only increase output torque, but to give slow spindle speed; a spindle rotation speed of 20-30 RPM is good. A 230-volt reversible motor with integral gearing to give 50 foot-pound torque at a speed of 75 RPM was originally tried. That proved inadequate even with the counterbalancing system. A further gear reduction of 4:1 using a chain drive proved acceptable. That was with a 65 pound counter-balance weight, and a 200 pound far-end weight. A surplus synchro connected to the gear motor drive at the closest-to-the-house side-tower permits remote indication of the configuration at the operating position during dark hours. It was necessary to use a gear step-down arrangement so that the synchro makes only one revolution of the entire configuration change.

acknowledgements

A project of this size cannot be accomplished without help. First I'd like to thank Marshall Etter, W2ER, whose long-distance correspondence gave me a fuller understanding of the practical aspects of rhombic antenna construction gleaned from his many years at RCA Riverhead Receiving Station; his knowledge boosted my confidence as well. Fred McGinnis, WD4KJZ, assisted in handling the lines and furnished muscle as we erected four towers — two 70-footers, one 80-footer, and 70 feet of a 100-footer. (Fred is 70+ years old and I was 63 at the time, so we made a great team.) John Fleming, WD4FFX, put up the last 30 feet of the 100 footer, which I declined to climb. Bill McCune, W2IRC, built the spindles and other mechanical parts and provided assistance and advice in the guying of the towers. Richard Bluhm, W2KXD, gave counsel on the Foster charts and design work. Alan Sielke, a non-ham, gave advice on structural loads on the towers from his civil engineering background and loaned me the transit. Norman Gertz, K1AA, furnished the 600-ohm terminating resistor as well as old military publications pertaining to rhombics. Gene Black, W2LL, furnished the old-style 600-ohm DPDT antenna relays needed for direction reversal. Millie Elwell, KA4ECM, helped in the initial survey of the rhombic towers and contributed encouragement and patience.

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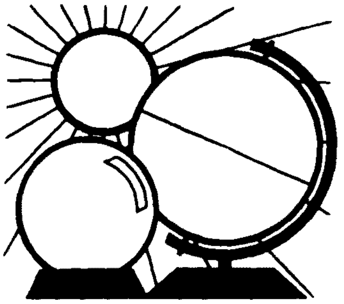

DX FORECASTER

Garth Stonehocker, K0RYW

vernal equinox DX

The average solar radio flux is often higher in March than it is in any other month of the year. During the first three months of the year the earth is at perigee with the sun; the long winter nights allow more time — after the earth's daily production of ions — for upward drift and the diffusion of ions into the F2 region. The F2 region contains the maximum ion density (foF2), which usually defines the maximum usable frequency (MUF) for DX paths.

Throughout these winter months, the foF2, the major variable factor in calculating MUF (MUF = foF2•factor), is accumulated a little at a time, day by day, and the highest monthly average of the year usually occurs during this quarter. Even during the year of the sunspot minimum, when the solar flux variation is small, an enchanced F2 region can be expected to build up in winter.

Geomagnetic storms during these months, however, may disrupt our midlatitude ionosphere. During the equinoctial months the earth's (dipole) magnetic field is sufficiently perturbed by solar wind particles flowing into the auroral zone at 50-70 degrees north geographic latitude to cause the midlatitude ionosphere to be depleted. Below the auroral zone, the ionosphere develops a trough that extends southward, mainly on the dark side of the earth (i.e., at night) for two to three days in a row. Only near the equator (between ±20 degrees *geomagnetic* latitude) do the geomagnetic disturbances enhance ionization; this is the reason for the higher MUF and ionospheric tilts that give rise to transequatorial (T.E. or one-long hop) propagation. This T.E. is characteristic of the equinoctial months — in the spring more than in autumn — and throughout winter in general.

springtime QRN

March and April are months in which spring storms bring rain to much of the northern hemisphere. Fronts of warm and cold air generate the first major thunderstorms of the year, producing static that reduces the signal-to-noise ratio of received signals, thereby lowering readability.

The cumulative effect of thunderstorm static worldwide is the main cause of high noise levels on the lower frequency HF bands. However, as a storm front approaches your area, a significant increase in the noise level is heard. One first notices this increase at a one-hop distance away (about 600 to 1200 miles or 960 to 1920 km) when the storm front is about one day west of your location. The noise level usually decreases after that until the storm reaches within a ground-wave's distance (50 to 60 miles or 80 to 96 km). Individual discharges can be heard. As the storm draws nearer, its sounds become part of the "local noise;" as it moves away, its noise decreases, then increases again as the front reaches the one-hop distance point a day or so later. (You can correlate this with storm progress reports on local television.) You can save time in looking for rare DX by tracking storms in order to pinpoint when the most favorable listening conditions are likely to occur.

last minute-forecast

T.E. can be expected on the higher frequency bands (10 through 30 meters) during the first two weeks of March. The effect will not be as pronounced as it was last year, but should still provide good transequatorial openings on the higher of these bands. Look for the best openings to occur when the geomagnetic field is disturbed (high A and K figures) toward the end of the second week. The rest of the month will be better for low band night time DX operation. In terms of QRN buildup, this month will be one of the last quiet ones, but only between storms. Spring equinox occurs on March 20th at 1024 UTC. The moon is full on the 17th and at perigee on the 16th.

band-by-band summary

Ten, fifteen, and twenty meters will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of the bands will be shorter and will occur closer to local noon. Transequatorial propagation on these bands will be more likely toward evening during conditions of high solar flux and a disturbed geomagetic field.

Thirty and forty meters will be useful almost 24 hours a day. Daytime conditions will resemble those on 20 meters, but skip and signal strength may decrease during midday on days coinciding with high solar flux values. Nighttime use will be good except after days of very high MUF conditions. Generally the usable distance is expected to be somewhat greater than that achieved on 80 at night.

Eighty and one-sixty meters, the nighttime DXer's bands, open just before sunset and lasts until the sun comes up on the path of interest. Except for daytime short-skip signal strengths, high solar flux values have little effect. Geomagnetic disturbances more evident near equinox cause signal attenuation and fading on polar paths. Noise will be very noticeable on these lower frequency bands.

ham radio

WESTERN USA

GMT	PST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
0000	4:00	20	20	15	10	15	10	10	20
0100	5:00	20	20	15	10	15	10	10	20
0200	5:00	20	20	20	10	15	10	10	20
0300	7:00	20	20	20	10	15	10	10	20
0400	8:00	20	30	20	10	15	15	10	20
0500	9:00	20	30	20	10	20	15	15	20
0000	10:00	20	30	20	15	20	15	15	20
0700	11:00	20	30	20	15	20	20	15	20
0800	12:00	*20*	*30*	20	15	20	20	20*	*20*
0900	1:00	*20*	*30*	20	15	20	*20*	*20*	*30*
1000	2:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1100	3:00	*30*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1200	4:00	*30*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1300	5:00	*30*	20	*20*	*20*	*20*	*20*	*20*	*30*
1400	6:00	*30*	20	15	*20*	*20*	*20*	*20*	*30*
1500	7:00	*30*	20	15	*20*	*20*	20	*20*	*30*
1000	8:00	*30*	20	15	*20*	*20*	15	*20*	*30*
1700	9:00	*30*	20	15	20	20	15	20	30
1800	10:00	30	20	15	20	20	15	15	30
1900	11:00	30	20	10	15	15	15	15	20
2000	12:00	30	20	10	15	15	10	15	20
2100	1:00	30	20	10	15	15	10	15	20
2200	2:00	20	20	10	15	15	10	15*	20
2300	3:00	20	20	15	10	15	10	10	20

MARCH

MID USA

MST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
5:00	30	20	15	15	15	10	10	20
6:00	30	20	15	15	15	10	10	20
7:00	30	20	20	20	15	10	10	20
8:00	30	20	20	20	15	10	10	20
9:00	30	30	20	20	15	15	15	20
10:00	*30*	30	20	20	20	15	15	20
11:00	*30*	30	20	20	20	15	15	20
12:00	*30*	30	20	20	20	20	20	20
1:00	*30*	*30*	20	*20*	20	20	20	*30*
2:00	*30*	*30*	*20*	*20*	*20*	*20*	*20*	*30*
3:00	*30*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
4:00	*20*	*30*	*20*	*15*	*30*	*20*	*20*	*30*
5:00	*20*	*30*	*20*	*15*	*30*	*20*	*20*	*30*
6:00	*20*	30	15	15	*20*	*20*	*20*	*30*
7:00	20	20	15	15	*20*	*20*	*20*	*30*
8:00	20	20	15	10	*20*	20	*20*	*30*
9:00	20	20	15	10	*20*	15	*20*	*30*
10:00	20	20	15	10	20	15	20	*30*
11:00	20	20	10	10	20	15	15	30
12:00	20	20	10	10	15	15	15	20
1:00	20	20	10	10	15	10	15	20
2:00	20	20	10	10	15	10	15	20
3:00	20	20	10	10	15	10	15*	20
4:00	30	20	15	15	15	10	10	20

EASTERN USA

CST	EST	N (ASIA FAR EAST)	NE (EUROPE)	E (S. AFRICA)	SE (CARIBBEAN S. AMERICA)	S (ANTARCTICA)	SW (NEW ZEALAND)	W (OCEANIA AUSTRALIA)	NW (JAPAN)
6:00	7:00	30	20	15	15	15	10	10	20
7:00	8:00	30	20	15	15	15	10	10	20
8:00	9:00	30	20	20	20	15	15	10	20
9:00	10:00	30	30	20	20	15	15	10	20
10:00	11:00	30	30	20	20	20	15	15	20
11:00	12:00	*30*	30	20	20	20	15	15	20
12:00	1:00	*30*	30	20	20	20	20	15	20
1:00	2:00	*30*	*30*	20	20	20	20	20	20
2:00	3:00	*30*	*30*	20	*20*	20	20	20	*30*
3:00	4:00	*30*	*40*	*20*	*20*	*20*	20	*20*	*30*
4:00	5:00	*20*	*40*	*20*	*20*	*30*	*20*	*20*	*30*
5:00	6:00	*20*	*30*	*20*	*15*	*30*	*20*	*20*	*30*
6:00	7:00	*20*	*30*	*15*	*15*	*20*	*20*	*20*	*30*
7:00	8:00	20	30	15	15	*20*	*20*	*20*	*30*
8:00	9:00	20	20	15	15	*20*	*20*	*20*	*30*
9:00	10:00	20	20	15	10	*20*	*20*	*20*	*30*
10:00	11:00	20	20	15	10	20	*15*	20	*30*
11:00	12:00	20	20	10	10	20	15	20	*30*
12:00	1:00	20	20	10	10	20	15	15	30
1:00	2:00	20	20	10	10	15	15	15	20
2:00	3:00	20	20	10	10	15	10	15	20
3:00	4:00	20	20	10	10	15	10	15	20
4:00	5:00	20	20	10	10	15	10	15*	20
5:00	6:00	30	20	15	15	15	10	10	20

The italicized numbers signify the bands to try *during* the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

VHF/UHF WORLD

Joe Reisert W1JR

keeping VHF/UHFers up-to-date

I'm often asked questions such as: "How do I know what's the latest VHF/UHF technique? Where can I meet other VHF/UHF'ers? How can I keep up to date on important VHF/UHF happenings?" The answers to questions like these are largely a function of interests and the bands used; and thus can vary considerably.

Because these questions are so frequently asked, I decided to answer them in this month's column. I'll try to cover as many areas as possible, as objectively as I can. If I neglect to mention a valuable VHF/UHF resource with which you're familiar, please let me know so we can share it with others in a later column.

periodicals

Probably the most popular source of information for VHF/UHF'ers is contained in the monthly Amateur Radio magazines (**table 1**).

ham radio has always been a reliable source of information on VHF/UHF design and techniques. Since January, 1984, this column has appeared monthly. In addition to various articles on VHF/UHF subjects (see Cumulative Index, December, 1984, page 128), the annual antenna issue (May), the annual VHF/UHF issue (which first appeared in July, 1984), and the annual receiver issue (November) are of particular interest to VHF/UHF'ers.

CQ has recently added a VHF/UHF column. It also features articles on VHF/UHF techniques and equipment. In addition it has announced that it will resume its VHF contesting program with a VHF WPX contest in July, 1985.

QST also publishes articles of interest to VHF/UHF'ers. The monthly columns "The World Above 50 MHz" by Bill Tynan, W3XO, and "The New Frontier" by Bob Atkins, KA1GT, cover current VHF/UHF standings and records and reports on VHF/UHF activity. ARRL-sponsored VHF/UHF, and EME contest results are also published regularly.

73 has no formal VHF/UHF column but publishes articles featuring VHF, UHF, and microwave techniques.

Several periodicals published outside the United States may also be of interest to stateside VHF/UHF'ers. The most notable are *CQ-DL*, *CQ Ham Radio*, *Ham Journal*, and *Radio Communication* (see table **2**).

CQ-DL, the official journal of the DARC (Deutscher Amateur Radio Club) of West Germany, is written entirely in German. Like *ham radio* and *QST*, it features a monthly VHF/UHF column as well as articles of interest to VHF/UHF'ers.

table 1. Major Amateur Radio periodicals published in the United States.

ham radio, Communications Technology, Inc., Greenville, New Hampshire 03048. Issued monthly. One-year subscription: $19.95 (U.S.A.).

CQ, CQ Publishing Company, 76 North Broadway, Hicksville, New York 11801. Issued monthly. One-year subscription: $16.00 (U.S.A.).

QST, ARRL, 225 Main Street, Newington, Connecticut 06111. Issued monthly. Annual membership in ARRL: $25.00 ([U.S.A.] includes subscription to QST).

73, 73 Subscription Department, Box 931, Farmingdale, New York 11737. Issued monthly. One-year subscription: $25.00 (U.S.A.).

CQ Ham Radio is the major monthly publication in Japan. Written in Japanese, it usually runs over 500 pages per issue. *Ham Journal*, also in Japanese, is a smaller, quarterly publication issued by the same publisher. Each issue of *Ham Journal* is based on a central theme. For instance, the Spring, 1984, issue was devoted almost entirely to EME.

Radio Communication ("*Radcom*") is the official journal of the RSGB (Radio Society of Great Britain). Like *QST*, it features articles on VHF/UHF and two VHF/UHF columns, "4-2-70" (70, 144, and 430 MHz) by Ken Willis, G8VR, and "Microwaves" by Mike Dixon, G3PFR. It is interesting to note

that British VHF/UHF'ers appear to have different interests than their American counterparts. Reading "*Rad Com*," the American reader senses a greater interest in portable operation and a large population of users on microwave frequencies such as 23, 13, and 3 cm. Unlike typical American designs, British designs tend to emphasize inexpensive or particularly clever design techniques.

professional periodicals

Several professional periodicals may also be of interest to VHF/UHF'ers (see **table 3**). In particular, they are the *IEEE Microwave Theory and Techniques* (*MTT*),* the *IEEE Transactions on Antennas and Propagation* (*PGAP*), *Microwave Journal*, *Microwaves & RF*, *Microwave Systems News*, and *RF Design*.

IEEE's *PGAP* and *MTT* are highly mathematical in nature but usually are devoted to the latest state-of-the-art developments in their respective fields of antennas, propagation, and microwaves. Only members of the IEEE may subscribe.

Microwave Journal, *Microwaves & RF*, *Microwave Systems News*, and *RF Design* are controlled-circulation publications available free of cost to "qualified" industry professionals and by subscription to "non-qualified" persons. These magazines specialize in reporting on the latest design techniques and equipment, with generous amounts of space devoted to advertisements and product reviews.

special interest publications

So far we've been talking mainly about magazines. There is, however, a substantial number of newsletters and small journals that are specifically written for VHF/UHF'ers. These are generally either "open distribution" publications or society or club publications.

The primary "general distribution" types are *VHF+ Trading Post*, *VHF/UHF and Above Information Exchange*, *2-meter EME Bulletin*, *220 Notes*, *432 and Above EME News*, *VHF/UHF Newsletter*, *DUBUS*, *VHF Communications*, *Amateur Television Magazine*, and *ORBIT* (see **table 4**).

VHF+ Trading Post is a new monthly newsletter that specializes in news and advertisements of interest to the VHF/UHF'er. *VHF/UHF and Above Information Exchange*, another new publication, with articles on VHF, UHF, and EME, as well as VHF/UHF'er comments and station descriptions. It also features K2UYH's "*432 and Above EME News*." The *2-Meter EME Bulletin* spotlights the latest happenings, news items, and station descriptions of interest to 2-meter EME'ers. All three publications stepped in to fill the gap when "*The Lunar Letter*" ceased publication.

220 Notes is dedicated to coverage of weak signal and FM/FM repeater activity on the 220-225 MHz band. Recent issues featured detailed background material on the Land Mobile industry's attempt to remove this band from the Amateur Radio service. *432 and Above EME News*, compiled by Allen Katz, K2UYH, features 70, 23, and 13-cm EME station reports, as well as some technical material and monthly EME schedules. A low-budget operation published mainly for the active EME'er on 70 cm and above, *432 and Above* cannot handle any more subscribers; it is, however, republished in the next available issue of *VHF/UHF and Above Information Exchange*.

The *RSGB VHF/UHF Newsletter* by David Butler, G4ASR, has just become available to readers outside of Europe. While it specializes in reporting on upcoming VHF and UHF activities in Europe, it also provides good material on meteor scatter, EME and new VHF and UHF techniques. *DUBUS* is a West German publication written mostly in English, but directed to European readers. *DUBUS* features information on stations, activities, and design for VHF/UHF operation.

VHF Communications is the English-language version of the German UKW Technik. Published quarterly, it features full-length articles on VHF, UHF, and microwave equipment design.

Amateur Television Magazine specializes in slow scan, fast scan, and satellite TV. It offers articles on equipment construction and operating techniques, and advertisements of interest to the video-oriented Amateur.

ORBIT, the journal of the Radio Amateur space program, is the official publication of AMSAT (Amateur Sat-

*Institute of Electrical and Electronics Engineers, Inc., 345 East 47th Street, New York, New York 10017.

table 2. Major Amateur Radio periodicals published outside the United States (contact publishers for U.S. subscription rates).

CQ-DL, Published by the German Amateur Radio Club (DARC), Postfach 1155, 3507 Baunatal 1, West Germany. Issued monthly.

CQ Ham Radio, CQ Publishing Company Ltd., 14-2 Sugamo 1-Chome, Toshima-Ku, Tokyo 170, Japan. Issued monthly.

Ham Journal, CQ Publishing Company, Ltd., 14-2 Sugamo 1- Chome, Toshima-Ku, Tokyo 170, Japan. Issued quarterly.

Radio Communication, RSGB Headquarters, Alma House, Cranborne Road, Potters Bar, Hertsfordshire, EN6 3JW, England. Issued monthly.

table 3. Major professional publications of interest to VHF/UHF'ers.

IEEE MTT, IEEE Service Center, 445 Hoes Road, Piscataway, New Jersey 08854. Issued monthly. IEEE membership required. Contact publisher for rates.

IEEE PGAP, IEEE Service Center, 445 Hoes Road, Piscataway, New Jersey 08854. Issued monthly. IEEE membership required. Contact publisher for rates.

Microwave Journal, Horizon House, 610 Washington Street, Dedham, Massachusetts 02026. Issued monthly. One-year subscription: $36.00.

Microwaves and RF, Hayden Publishing Company, Box 1419, Riverton, New Jersey 08077. Issued monthly. One-year subscription: $30.00.

Microwave Systems News, E.W. Communications, Inc., Box 50249, Palo Alto, California 94303-0249. Issued monthly. One-year subscription: $35.00.

RF Design, Cardiff Publishing Company, 1 East First Street, Duluth, Minnesota 55802. Issued bi-monthly. One-year subscription: $15.00.

table 4. Major VHF and UHF newsletters and publications.

VHF+ Trading Post, c/o Jack Parker, KC0W, P.O. Box 11023, Reno, Nevada 89510. Issued monthly. One-year subscription: $5.00.

VHF/UHF and Above Information Exchange, c/o Rusty Landes, KA0HPK, P.O. Box 270, West Terre Haute, Indiana 47885. Issued monthly. One-year subscription: $15.00.

2-Meter EME Bulletin, c/o Gene Shea, KB7Q, 417 Stadhauer, Bozeman, Montana 59715. Issued monthly. One-year subscription: $12.00.

220 Notes, c/o Walt Altus, WD9GCR, 215 Villa Road, Steamwood, Illinois 60103. Six issues per year. One-year subscription: $5.00

432 and Above EME News, c/o Allen Katz, K2UYH, 326 Old Trenton Road, RD 4, Trenton, New Jersey 08691. Issued monthly. (See text).

VHF/UHF Newsletter, RSGB Headquarters, Alma House, Cranborne Road, Potters Bar, Hertsfordshire EN6 3JW, England. Issued monthly. Contact publisher for rates.

DUBUS, c/o Claus Neie, DL7QY, D-7181 Rudolfsberg 24, West Germany. Issued quarterly. Contact publisher for rates.

VHF Communications, c/o Terry D. Bitten, Jahnstrasse 14, Postfach 80, D-8523 Baiersdorf, West Germany. Issued quarterly. Contact publisher for rates.

Amateur Television Magazine, c/o QCD Publications Inc., P.O. Box H, Lowden, Iowa 52255. Issued monthly. One-year subscription: $10.00.

ellite Corporation). It features construction articles, equipment reviews, station descriptions, and the latest information on Amateur Radio satellites.

club newsletters

There are numerous club publications often available to nonmembers at a nominal cost. Those that come to mind are *Six Shooter*, The *Sidewinders on Two Bulletin*, *The Southeastern VHF Society and 70-cm Net Newsletter*, *Northeast VHF News*, *Texas VHF-FM Society News*, *The West Coast VHF'er*, *Radiosporting*, and *Cheese Bits* (see **table 5**).

Six Shooter, the official journal of SMIRK (Six Meter International Radio Klub), specializes in coverage of 6-meter activities. You can join SMIRK *after* contacting six members on 6 meters. SMIRK also sponsors awards and an annual contest.

The *Sidewinders on Two Bulletin* is the official newsletter of SWOT (Sidewinders on Two), which specializes in 2-meter activity. Members are admitted after contacting 2 SWOT members on 2 meters. SWOT also sponsors an annual contest and awards.

The *Southeastern VHF Society and 70-cm Net Newsletter* is a publication of the Southeastern 70-cm net. In addition to 70-cm news, it features items of general and technical interest to VHF/UHF'ers.

Northeast VHF News, sponsored by the Northeast VHF Association, carries news of interest to local, national, and international VHF/UHF'ers. *Texas VHF-FM Society News* specializes in FM and FM repeaters in the Texas area as well as meeting notes from the Texas VHF-FM Society.

The West Coast VHF'er concentrates on information of interest to the California VHF/UHF'er. *Radiosporting* is a new magazine which promises to focus on the art of Amateur contesting. The Mount Airy VHF Radio Club, Inc. Pack Rats' *Cheese Bits* is a club newsletter that includes product reviews, an activity calendar, a swap and shop column, and short technical articles.

Let us not forget those publications that while no longer available, are still an excellent source of information if you can borrow copies. The first ones that come to mind are *The VHFER* (K7AAD), *6 UP* (*73*), *220 MHz EME Newsletter* (K5FF), *Northern California 220 News* (WA6GYD), *432 Bulletin* (W6FZJ), and *The Lunar Letter* (KI7D).

Other short publications or newsletters while general in nature, often print information of interest to VHF/UHF'ers (see table **6**). They are *The ARRL Letter*, *World Radio*, *QEX*, *Westlink Report* (formerly HR Report), *The W5YI Report*, *Gateway*, and *DX Bulletin*.

The *ARRL Letter* is primarily dedicated to keeping the Amateur informed of Amateur Radio news in general. As such, it provides up-to-date information in a timely fashion. *World Radio* based in Sacramento, California, is an Amateur Radio newspaper. It publishes articles of general interest to the Amateur and features several columns including ones on DX, antennas, the FCC, and OSCAR. *QEX*, the ARRL experimenters' exchange, is a monthly ARRL newsletter specializing in articles that may be too technical in nature for

table 5. VHF/UHF club and organization bulletins.

Six Shooter, c/o Ray Clark, K5ZMS, 7158 Stone Fence, San Antonio, Texas 78227. Issued quarterly, Annual membership dues $6.00, plus $3.00 for subscription.

The Sidewinders on Two Bulletin, c/o Harry Arsenault, K1PLR, 603 Powell Avenue, Erie, Pennsylvania 16505. Issued monthly, One-year subscription: $10.00.

Southeastern VHF Society and 70-cm Net Newsletter, c/o Charles Osborne, WD4MBK, 131 Saratoga Drive, Lawrenceville, Georgia 30245. Issued quarterly. One-year subscription: $5.00.

Northeast VHF News, c/o Lewis Collins, W1GXT, 10 Marshall Terrace, Wayland, Massachusetts 01778. Six issues per year. One-year subscription: $3.00.

TX VHF-FM Society, c/o Robert McWhorter, K5PFE, Box 461, Jasper, Texas 75951. Issued bi-monthly. One-year subscription: $6.00.

The West Coast VHFer, 560 West Yucca Street, Oxnard, California 93033. Issued monthly. One-year subscription: $10.00.

Radiosporting, c/o Yuri Blanarovich, VE3BMV, Box 65, Don Mills, Ontario, M3C 2R6, Canada. Issued monthly. $12.00 annual membership *plus* $16.00 for one-year subscription.

The Pack Rats' Cheese Bits, c/o Doc Cutler, K3GAS, 7815 New Second Street, Elkins Park, Pennsylvania 19117. Issued monthly. One-year subscription: $2.50.

table 6. Newsletters of a general nature that often cover news of interest to VHF/UHF'ers.

The ARRL Letter, c/o ARRL, 225 Main Street, Newington, Connecticut 06111. Issued bi-weekly. One-year subscription: $19.50. (ARRL members only).
QEX, c/o ARRL, 225 Main Street, Newington, Connecticut 06111. Issued monthly. One-year subscription: $6.00 (ARRL members only).
Westlink Report, c/o Poco Press, 11119 Allegheny Street, Sun Valley, California 91352. Issued bi-weekly. One-year subscription: $22.50.
The W5YI Report, Box 10101, Dallas, Texas 75207. Issued bi-weekly. One-year subscription: $24.00.
Gateway, c/o ARRL, 225 Main Street, Newington, Connecticut 06111. Issued bi-weekly. One-year subscription: $6.00 (ARRL members only).
World Radio, 2120 28th Street, Sacramento, California 95813. Issued monthly. One-year subscription: $10.00.

QST. Articles on experimental aspects of VHF/UHF are always welcome. Geoffrey Krauss, WA2GFP, writes a bi-monthly column for *QEX*.

The two major Amateur Radio bi-weekly newsletters are *The Westlink Report* (formerly *HR Report*) and *W5YI Report*. They primarily cover Amateur Radio news items, with emphasis on FCC matters, legislative problems, recent or upcoming events. VHF/UHF news is printed as it becomes available. *Gateway*, the latest ARRL publication, specializes in packet radio communications in a format similar to *QEX*. *Gateway* should be especially interesting to those people using packet radio for VHF meteor scatter.

DX Bulletins

Many DX bulletins are published regularly. These newsletters usually provide important propagation information as well as VHF/UHF reports (when received). This category includes *The DX Bulletin* (K1TN), *QRZ DX* (W5KNE), *LIDXA Bulletin* (W2IYX), and The *DX'ers Magazine* (W4BPD), among others. *The Northern California DX Foundation Newsletter* (N6ST) has contributed to at least two VHF/UHF DXpeditions.

catalogs and advertisements

Although advertisements and catalogs are intended to sell products, they can also provide useful information about the state-of-the-art, availability of components and similar subjects. Don't overlook advertisements as a source of helpful information. New product reports and reviews in periodicals and newsletters can also be useful.

Amateur reference materials

Numerous books and references useful to the VHF/UHF'er include but are not limited to *The ARRL 1985 Handbook for The Radio Amateur, Radio Handbook, VHF/UHF Manual, VHF Handbook for Radio Amateurs, The Radio Amateur's VHF Manual, VHF for the Radio Amateur, The UHF Compendium* (Parts 1 and 2), *The Satellite Experimenter's Handbook, ARRL Antenna Book, The Microwave Newsletter Technical Collection*, and *From Beverages Thru OSCAR — A Bibliography, with Addendum* (see **table 7**).

The ARRL 1985 Handbook for The Radio Amateur and Bill Orr's *Radio Handbook* each have several chapters devoted to VHF/UHF communications. The *VHF/UHF Manual* (RGSB), *VHF Handbook for Radio Amateurs, The Radio Amateur's VHF Manual*, and *VHF for the Radio Amateur* are entirely devoted to VHF/UHF and Microwaves, with information on antennas, receivers, transmitters, etc. The *Microwave Newsletter* may be out of print and only a few dozen copies of the *Bibliography* are still available from the source. Both include basic material useful to VHF/UHF'ers.

The *UHF Compendium* (Parts 1 and 2), a translation of a German publication, is a great source of information for designing receivers, antennas, and transmitters. Heavy emphasis is placed on practice and construction. *The Satellite Experimenter's Handbook* is a history of and guide to all of the Amateur Radio Satellites. It shows readers how to find the various satellites and recommends specific antennas, preamplifiers, and transmitters. The *ARRL Antenna Book* is an excellent reference for both HF and VHF. *The Microwave Newsletter Technical Collection* consists of edited extracts from the best UHF and microwave notes published in the RSGB *Microwave Newsletter* since 1980.

table 7. Recommended VHF, UHF, and Microwave references.

The ARRL 1985 Handbook for the Radio Amateur, ARRL, 225 Main Street, Newington, Connecticut 06111, $15.00.[1]
Radio Handbook, William I. Orr, W6SAI, $12.95.[1]
VHF/UHF Manual, Pat Jessop, G5JP. Published by the RSGB. $17.50.[1]
VHF Handbook for Radio Amateurs, Herb Brier, W9EGQ, and William I. Orr, W6SAI, $11.95.[1]
The Radio Amateur's VHF Manual, Edward P. Tilton, W1HDQ, ARRL, 225 Main Street, Newington, Connecticut 06111. (out of print)
VHF for the Radio Amateur, Frank C. Jones, W6AJF, CO Publications, 76 North Broadway, Hicksville, New York 11801 (out of print).
The UHF Compendium, Parts 1 and 2, K. Weiner, DJ6HO, Editor (out of print).
The Satellite Experimenter's Handbook, Martin Davidoff, K2UBC, ARRL, 225 Main Street, Newington, Connecticut 06111. $10.00.[1]
The ARRL Antenna Book, 14th edition, Gerald L. Hall, K1TD, Editor, $8.00.[1]
The Microwave Newsletter Technical Collection, J. Gannaway, G3YGF, and S.J. Davies, G4KNZ. Published by the RSGB. $10.00.[1]
From Beverages Thru OSCAR — A Bibliography, (with Addendum). *Volume 1* (1945-1978) 620 pp, $29.95, *Volume 2*, (1979-1981) 144 pp, $9.95. Both volumes: $35.95. Contact author, Rich Rosen, K2RR, at *ham radio*.

Note 1. Available from Ham Radio's Bookstore, Greenville, New Hampshire 03048. *Add $3.50 for shipping and handling.*

table 8. Calling frequencies.

6 meters	50.110 MHz
2 meters	144.200 MHz
135 cm	220.100 MHz
70 cm	432.100 MHz
23 cm	1296.100 MHz
13 cm	2304.100 MHz

From Beverages thru OSCAR — A Bibliography is a unique reference book that lists, in chronological order, virtually every Amateur Radio article written in the last 30 or so years through 1981.* Over 36,000 articles appearing in the Amateur press (as well as in some professional publications) are included. As such, it is a good source for determining what resources are available, where and when they were published. All of these valuable references are a must for the library of any well-rounded, informed VHF/UHF'er.

nets

Nets are a valuable and timely source of information for those interested in VHF, UHF, and microwaves. Local and VHF/UHF club nets are quite common. SMIRK and SWOT sponsor numerous 6 and 2-meter nets; their members can direct you to nets in your local area.

There are several HF nets or "hangout" frequencies where VHF/UHF'ers can often be found. The 6-meter enthusiasts can often be found on 28.885 MHz when 10 and/or 6 meters is open. Much crossband and 6-meter DX was coordinated on this frequency when F2 propagation was common at the peak of solar cycle 21.

The CSVHF (Central States VHF Society) has a net on 3818 kHz, usually at 9:30 PM CST on Sunday evenings. VHF/UHF'ers, especially those in the central United States, often use this frequency at night — and especially during meteor showers — to exchange information and schedules.

14,345 kHz has long been used by Europeans as a VHF scheduling frequency. Every Saturday and Sunday the 70-cm EME net meets on 14,345 kHz at 1600 UTC. The 2-meter EME net follows at approximately 1700 UTC. These nets, although primarily devoted to EME, are a good source of information on important VHF/UHF happenings. Many VHF'ers are known to monitor (but don't necessarily check in) these nets to get the latest "scoop." The EME'ers and many VHF/UHF'ers also use OSCAR 10 as an intercom, and can often be heard on the downlink frequency of 145.950.

Other HF gatherings also take place. Of particular note is a large group of VHF'ers that seems to "hang out" on 160 meters where they swap stories or lies. I guess the 160-meter crowd has something in common with VHF/UHF'ers: *they also enjoy suffering with weak signals and noise*!

Don't forget the weekly *OSCAR* nets (see *QST* for frequencies, dates, and times). They are a good source of VHF/UHF as well as satellite information.

calling frequencies

Often the VHF or UHF frequencies are quiet because there are simply fewer VHF/UHF'ers than HF'ers. VHF/UHF propagation is also more restricted than HF. As a result, the VHF'ers have established "calling frequencies," where everyone can monitor, call CQ occasionally, or establish communications with someone else who monitors that same frequency. It is common courtesy to slide off the calling frequency after making contact so that someone else can use it.

Monitoring calling frequencies can be a good way to find out who's active and what's happening (see **table 8**). The prime calling frequencies in the USA are 50.110, 144.2, 220.1, 432.1, and 1296.1 MHz. As previously mentioned, the OSCAR 10 downlink frequency of 145.950 MHz also serves as a gathering and calling frequency for VHF/UHF'ers.

activity

The Northeast VHF Association has introduced a popular system concentrating activity and thereby increasing the likelihood of meeting other interested VHF/UHF'ers. This system is built around "activity nights" and "activity hours," see **tables 9** and **10**. Basically, the system works this way. Sunday night is 6-meter activity night; Monday is 2 meters; Tuesday is 135 cm; Wednesday is 70 cm; Thursday is 23 cm. (Anyone for 13 cm on Fridays?) By concentrating on a single band each night, VHF/UHF'ers, especially those who operate multiple bands, are more likely to find other interested parties. These nights are a great source of information exchange. During VHF/UHF contests it may be difficult to catch all active stations, especially those that operate several bands. Again, a technique recently suggested by the Northeast VHF Society consists of concentrating afternoon contest activity by establishing specific activity hours. 135-cm activity starts at 2 PM, 70 cm at 3 PM, and 23 cm at 4 PM. For many years a similar system has been used in the mornings and evenings, with 2 meters at 7 AM/PM, 135 cm at 8 AM/PM, etc. When everyone adheres to this plan, there is less likelihood of missing other active stations. It's also a good way to just keep in touch.

contests, conferences and shows

Contests can be good sources of information exchange. The ARRL sponsors VHF, UHF, EME, and SPRINT (single band short duration contest) contests throughout the year. (I try to list all contests at the end of this column every month). So do SMIRK (in June) and SWOT (in July). The Europeans also have many contests, with the biggest being the International VHF Contest held on the first weekend of October.

***From Beverages Thru OSCAR — A Bibliography** is in the process of being updated to include all published articles up to, and including, December, 1984.

table 9. Suggested VHF/UHF "activity nights."

6 meters	Sunday
2 meters	Monday
135 cm	Tuesday
70 cm	Wednesday
23 cm	Thursday
13 cm	Friday

table 10. Suggested "activity hours" during VHF/UHF contests.

2 meters	7AM	1PM	7PM
135 cm	8AM	2PM	8PM
70 cm	9AM	3PM	9PM
23 cm	10AM	4PM	10PM
13 cm	11AM	5PM	11PM

One excellent way to stay informed and meet with other interested VHF, UHF, and Microwave enthusiasts is to attend conferences. Often antenna and noise figure measurements are conducted; this can be an excellent way to see how your receiver, preamplifier, or converter and antenna stack up against others. At the same time, one can swap circuits and tips on improving performance, see firsthand the latest state-of-the-art tricks and devices.

The Dayton Hamvention is one of the first conferences held each year. Recently the Hamvention has sponsored a VHF/UHF program under the guidance of WA8ONQ and has now added noise figure and antenna measurements. The long-standing West Coast VHF Conference, held in California in early May, the Eastern VHF/UHF Conference, held in mid-May in New Hampshire, and the CSVHF Society Conference held the last weekend of July in the midwest, are totally dedicated to technical sessions on VHF/UHF and microwave techniques. All of these conferences sponsor antenna and noise figure measurements as well.

The Mid-Atlantic States Conference, held the first weekend in October in eastern Pennsylvania includes technical talks and a flea market. For those who travel abroad, Europeans have similar conferences in both Germany and England (contact the DARC or RSGB). A new conference dedicated solely to 1296 and 2304 MHz will be held in Colorado this September. Notes and handouts are frequently one of the highlights of these conferences. Recently the CSVHF Society began issuing a set of proceedings at their conference (contact KØDAS).

Finally, many area and ARRL conferences (especially the ARRL National) often feature VHF and UHF programs. An excellent source of information, these programs provide ample opportunity to meet and talk with experts in the field. As in the past, I will try to announce these conferences and the names of appropriate contact people at the end of each month's column.

awards programs

Awards programs generate activity and challenge VHF/UHF and microwave enthusiasts to improve their gear and operating techniques. The most common awards are the IARU WAC, the ARRL WAS, and the ARRL VUCC. Information on these awards can be obtained by sending an SASE to ARRL. The RSGB has members-only Microwave Transmitting Awards somewhat like the VUCC but specializing on the 23 cm and higher bands. Contact the RGSB for further details. The VUCC Award, which has done much to stimulate activity on the VHF/UHF bands, is based on contacts with different grid squares, a section of land included within a block measuring 2 degrees wide (longitude) by 1 degree high (latitude).[1] The WAS and VUCC standings boxes in "The World Above 50 MHz" in *QST* are an excellent way to discover who's active in your area.

conclusion

VHF and UHF'ers have a relatively poor track record when it comes to "Communicating." Exchange of information — especially in a timely way — has always been a problem. In this column therefore, I've tried to identify the principal sources of information for VHF/UHF'ers and HF'ers alike. I hope this material will provide readers with new sources of information; if you'll let me know of any important material I may have missed, I'll share it with *ham radio*'s readers in a later column.

reference

1. John Lindholm, W1XX, "VHF/UHF Century Club Awards," *QST*, January, 1983, page 49.

upcoming VHF/UHF events

March 21: *Optimum date for TE contacts.*

September 20-22: *I've just been informed by Don Hillard, WØPW, Box 563, Boulder, Colorado 80306, that he intends to sponsor a 1296 and 2304 MHz conference in Estes Park, Colorado. Drop Don an SASE for additional information on this gathering.*

ham radio

NEW products

new ICOM headquarters

ICOM America's new 40,000 square foot corporate headquarters/sales and service center in Bellevue, Washington, has been completed. Growth in its expanding line of Amateur Radio equipment had caused ICOM America to outgrow its prior facility; the new building allows for expanded service and warehouse areas as well as areas designated for future engineering and manufacturing divisions.

For information on ICOM products, contact ICOM America, Inc. at its new address: 2380-116 Avenue NE, Bellevue, Washington 98004.

Circle #301 on Reader Service Card.

handheld analog/digital multimeters

John Fluke Manufacturing has introduced two new heavy-duty analog/digital multimeters — the Fluke 25 and the Fluke 27 — designed specifically for industrial use. The new Fluke 20 Series combines the accuracy of a digital meter with the dynamic measurement capabilities of an analog meter. Built to endure environmental and electrical abuse, these sealed meters can withstand drops, shock, vibration, contaminants, moisture, and other harsh conditions.

Both are available in either safety yellow or dark charcoal gray. Their liquid crystal display works even at extreme temperatures; operation is guaranteed from −15 degrees C to 55 degrees C and to −40 degrees C for 20 minutes. Typical continuous operation is from −20 degrees C to 60 degrees C.

The Fluke 25 is priced at $229; the Fluke 27 at $259.

For complete details, contact John Fluke Mfg. Co., Inc., P.O. Box C9090, Everett, Washington 98206.

automatic antenna tuner

The Heath Company has expanded its Amateur Radio line to include the new SA-2500 Auto-Tune Antenna Tuner. The SA-2500 features an efficient, continuously variable roller inductor that can be preset for 18 different frequencies.

The SA-2500 permits the user to preset high and low frequencies on each of the nine bands from 160 to 10 meters. In the Auto mode this tuner sets the roller inductor to the preselected value and automatically adjusts the preset for a proper match. A remote capability allows selected frequencies to be automatically tuned to the proper SWR using only transmitter band switches, provided the transmitter is equipped for remote operation.

Manual tuning is simplified with three front panel lever switches and dual wattmeters. The wattmeters read forward and reflected average power and SWR in two ranges. An auto-range circuit automatically switches the wattmeters to the appropriate range.

The SA-2500 effectively tunes and matches unbalanced feed lines and single-wire antennas at the full legal power limit of a station. The SA-2500-1 4:1 Balun Accessory can be added for use with balanced ladder line antennas. A front panel coax switch allows the user to select easily from three different, permanently connected antennas and bypass.

Heath's Auto-Tune Antenna Tuner installs directly into the transmission line. The internal SWR wattmeter bridge will measure power on all frequencies between 1.8 and 30 MHz, 200/2000 watts in the forward direction and 50/500 watts reflected. SWR readings on the reflected meter provide direct readings from 1:1 to 3:1.

For more information and a free catalog, contact Heath Company, Department 150-395, Benton Harbor, Michigan 49022.

Circle #303 on Reader Service Card.

"dishpositioner"

Electroaids Inc. has announced the introduction of its Electro-Scan '85 "dishpositioner." The reliability of the new Electro-Scan '85 has been increased by the inclusion of minor improvements in the state-of-the-art circuitry. The exterior of the control box has also been redesigned for greater attractiveness and a more contemporary appearance.

The Electro-Scan '85 uses an analog micro system and features a dial control that shows dish location by actual satellite name. The "dishpositioner" uses a 36-volt DC motor drive for safety, and offers precise positioning along with a lock and key on the control box for owner control.

Information on the unit is available from Electro-Com, Suite 112, 8459 N. Main Street, Dayton, Ohio 45415.

Circle #302 on Reader Service Card.

SHORT CIRCUIT HOTLINE

Building a current *ham radio* project? Call the Short Circuit Hotline any time between 9 AM and Noon, or 1 to 3 PM — Eastern time — *before* you begin construction. We'll let you know of any changes or corrections that should be made to the article describing your project.

(See "Publisher's Log," April, 1984, page 6, for details.)

short circuits

solar power

In the article, "Complete Solar Power for your ham station," by NH6N, (December, 1984, page 22, the note in **fig. 6** should be changed to read: "Adjust P-1 to have comparator go low at 0.5 A or greater as desired."

VIC-20-/ASR-33 printer

In the September, 1984 issue a line of coding was inadvertently omitted from the program listing in W2QLI's ham note, "VIC-20 printer" (page 88). **210 GET B$** should be inserted between lines 200 and 220. Other than that, both **figs. 1A** and **1B** are correct and the circuit/software combination should work.

The following additional information provided by the author should also aid in joining micros to teletype machines:

"One does not have to use the 9-pin terminal strip under the call box on the ASR-33. Socket No. 2 at the end of the call box on the ASR-33 has 15 pins (see **fig. 1**) and is easy to get at. These pins are connected to the terminal strip. Radio Shack sells a 12-pin plug. All one has to do is cut the top off the plug so that it will fit the 15-pin socket.

"I believe the interface will work with the Commodore 64 or any other micro, that has the necessary outputs and can accept the listing. (It should also be possible to use the ASR-35 and ASR-43.)

"I ran a big Centronics 101A printer using the **fig. 1B** interface. It required 2400 baud. (The ASR-33 provides 110 baud, ASCII.) I have other software for the VIC-20-to-ASR-33 such as "screen dump," "save on tape," and "run to the ASR-33."

The nice feature of this VIC-20-to-ASR-33 is that it works both ways. Type on the VIC-20, and it prints on the paper of the ASR-33; type on the ASR-33 keyboard, and it appears on both the screen and paper."

SHORT CIRCUIT HOTLINE

Building a current *ham radio* project? Call the Short Circuit Hotline any time between 9 AM and Noon, or 1 to 3 PM — Eastern time — *before* you begin construction. We'll let you know of any changes or corrections that should be made to the article describing your project.

(See "Publisher's Log," April, 1984, page 6, for details.)

603-878-1441

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Coming Events

ACTIVITIES

"Places to go..."

OHIO: Dayton Hamvention, April 26, 27, 28, Hara Arena and Exhibition Center, Dayton. Admission $8 advance, $10 at door. Good for all three days. Banquet $14 advance, $16 at door. Flea market space $17 in advance for all three days. Technical, ARRL and FCC forums. New products and exhibits. Special group meetings. YL forum. International VHF/UHF conference. Amateur of the Year Award. Special achievement awards. Pre-registration starts January 1, 1985. For further information; Dayton Amateur Radio Association, Box 44, Dayton, OH 45401 or phone (513) 433-7720.

NEW JERSEY: The Delaware Valley Radio Association's 13th annual Amateur Radio and computer equipment flea market, Sunday, March 17, 8 AM to 4 PM, New Jersey National Guard 112th Field Artillery Armory, Eggerts Crossing Road, Lawrence Township, Trenton. Advance registration $2.50, $3.00 at the door. Indoor and outdoor flea market area, dealers, and refreshments. Sellers bring own tables. Talk in on 146.52 and 146.07-.67 repeater. For tickets and space reservations: KB2ZY, Box 441B, RD#1, Stockton, NJ 08559. Please SASE.

ILLINOIS: LAMARSFEST '85 sponsored by the Libertyville and Mundelein Amateur Radio Society, Sunday, March 31, Lake County Fairgrounds, Grayslake. Doors open 8 AM. Set up 6 AM. Advance admission $2.00. $3.00 at the door. Indoor exhibits, code speed efficiency testing, free parking. Talk in on 146.94 simplex — 147.63-03 Waukegan Repeater. For information and reservations: LAMARS, Box 751, Libertyville, IL 60048.

ILLINOIS: The Sterling-Rockfalls Amateur Radio Society's Silver Anniversary Hamfest, March 10, Sterling High School Fieldhouse, 1608 4th Avenue, Sterling. Commercial distributors, dealers, free parking and a large flea market. Space for self-contained RV's overnight. Advance tickets $2.00. At door $3.00. Tables requiring electricity and all commercial tables $5.00. Others $3.00. For tickets, tables or information: Sue Peters, KA9GNR, PO Box 521, Sterling, IL 61081. (815) 625-9262. Talk in W9MEP 146.25/85.

OHIO: The Lake County Amateur Radio Association's seventh annual Lake County Hamfest and Computerfest, Sunday, March 31, Madison High School, Madison. 8 AM to 4 PM. Exhibitors 5:30 AM. Admission $3.00 advance and $3.50 at the door. Table and display space $5.00/6' table; $6.50/8' table. Plenty of free parking and all display space is indoors. Talk in on 147.81/.21. For information/reservations: SASE to Lake County Hamfest Committee, 713 W. Jackson, Painesville, Ohio 44077. (216) 952-9784.

INDIANA: The Indiana Hamfest (formerly the Martinsville Hamfest) sponsored by the Morgan County Repeater Association will be held March 10 at the Indiana State Fairgrounds Pavillion Building, Indianapolis. Admission $5.00 at the door. Premium table $40.00. Flea market table $8.00. Flea market space only $3.00. All tables by advance reservation only. Reserved table setup Saturday, March 9 from 3 to 9 PM. Space setup Sunday, March 10, 6 to 8 AM. Free parking. Talk in on 145.25. For table reservation or information SASE before March 1 to Aileen Scales, KC9YA, 3142 Market Place, Bloomington, IN 47401. (812) 339-4446.

PENNSYLVANIA: The third annual Southern Alleghenies Hamfest, sponsored by the Bedford, Altoona, Somerset, PA and Cumberland, MD Amateur Radio Clubs and Blue Knob Repeater Association. Sunday, April 14, 7 AM to 4 PM, Bedford County Fairgrounds. Admission $3.00. Tables $5.00 each. Tailgating $2.00. Dealers' setup Saturday, April 13. Talk in on Bedford Repeaters 145.49/89, 444.2 + 5 MHz and 146.52 simplex. For information: Joel Cunard, KB3TR, RD 6, Box 104, Bedford, PA 15522. (814) 623-9697.

MASSACHUSETTS: The Wellesley Amateur Radio Society's annual Spring Auction, Saturday, March 30, Wellesley Hills First Congregational Church, 207 Washington Street, Wellesley Hills. Check-in starts 10 AM. Auction 11 AM. Commission 15% with $1 minimum and $30 maximum. Food and drink available. No admission charge and plenty of free parking. Talk in on 147.63/03. For information: Nels Anderson, K1JR, (617) 872-5259.

ILLINOIS: Computer Central. Show and Swap. Sunday, March 3, Rand Park Field House, 2025 Dempster, Des Plaines. Information (312) 940-7547.

CALIFORNIA: The Tri-County ARA in cooperation with the Greater Los Angeles AR Group will conduct Amateur Radio exams in Pomona, Saturday, March 2. Pre-registration is required by February 16, 1985. Send completed FCC form 610, a copy of Amateur Radio license and SASE to T.C.A.R.A., PO Box 142, Pomona, CA 91769.

ILLINOIS: The 19th annual Rock River ARC Hamfest, Sunday, March 31, Lee County 4-H Center. Doors open 8 AM. Advance ticket donation $2.00. $3.00 at gate. 8' tables $5.00. Inside flea market space $3.00. Lunch will be served. Camping space available at a nominal charge. Talk in on 146.37/97 and 444.700/449.700. For tickets, tables, space or information: Shirley Webb, KA9HGZ, 618 Orchard St., Dixon, IL 61021 (815) 284-3811. Advance tickets available until March 15.

NEW JERSEY: The Split Rock Amateur Radio Association's annual Ham Auction, March 8, VFW Post 3401, Tabor Road, Rt. 53, Morris Plains. Doors open 7 AM. Auction starts 8 AM. Plenty of free parking. For information: PO Box 3, Whippany, NJ 07981 or K2RF Repeater 146.385/146.985.

OHIO: The Cincinnati QCWA Chapter 9's annual banquet in conjunction with the OOTC and the Dayton Hamvention, Friday, April 26, 7:30 PM, Neils Heritage House, 2189 S. Dixie Drive. For reservation and tickets ($12.50 pp) contact: Bob Dingle, Sec/Treas., 657 Dell Ridge Drive, Dayton, Ohio 45429. (513) 299-7114.

NEW JERSEY: The Chestnut Ridge Radio Club's Ham Radio flea market, Saturday, March 30, Education Building, Saddle River Reformed Church, East Saddle River Road and Weiss Road, Upper Saddle River. Tables $10.00 for first, $5.00 each additional. No admission fee. For information: Jack Meagher, W2EHD, (201) 768-8360 or Roger Soderman, KW2U (201) 666-2430.

NEW HAMPSHIRE: The Interstate Repeater Society's annual Flea Market, Saturday, March 16, Lions Club, Lions Avenue, Hudson. Talk in on 146.25/85 and 146.52 simplex. For table reservations: Interstate Repeater Society, PO Box 693, Derry, NH 03038. Or call Dick, WB8YGR (603) 889-3479.

NORTH DAKOTA: The Red River Valley and N.D.S.U. ARC's present "Hobbie Hi Tech 85", a Ham Radio/Computer show and swap meet, March 30, 8 to 5, Army National Guard Armory at Hector Field, Fargo. Reserved tables: Commercial $20; non-commercial — full $5.00, 1/2 $3.00. FCC exams by pre-registration. Talk in on 16/76. For information: Tim Gooding, WD0GUR, Event Chairman, 1006 Sheyenne Street, West Fargo, ND or call (701) 282-6630.

MASSACHUSETTS: 19/79 ARA of Chelsea will hold its annual flea market, Sunday, March 24, 11 AM to 3 PM, Ryan Hall at Ireson Building, 493 Western Avenue (Route 107) Lynn. Sellers setup 10 AM no admission charge. General admission $1.00. Table $6.00 advance or $8.00 at the door. For table reservations send checks to 19/79 ARA, PO Box 171, Chelsea, MA 02150.

19/79 ARA is sponsoring all level FCC exams, Saturday, March 16, Ryan Hall, Ireson Building, 493 Western Avenue (Route 107) Lynn, MA, 10 AM to 1 PM. For General, Advanced and Extra send form 610 with check for $4.00 payable to ARRL/VEC, 30 days prior to exam to Ralph Gandolfo, KA1E, 18 Murdock Drive, Peabody, MA 01960. For Tech, same requirements but send check and form 610 to Bob Kalustian, WA1DVR, 36 Columbia Road, Arlington, MA 02174. Novice there is no fee and walk-ins will be accepted but prefer prior registration. Send to Bob Kalustian at above address.

NEW JERSEY: The Shore Points ARC invites everyone to Springfest '85, Saturday, March 30, Atlantic County 4-H Center, Egg Harbor City, about 15 miles west of Atlantic City, 9 AM to 2 PM.

OPERATING EVENTS
"Things to do..."

MARCH 23-25: B.A.R.T.G. Spring RTTY Contest, 0200 GMT Saturday, March 23 to 0200 GMT Monday, March 25. Total contest period is 46 hours but no more than 30 hours of operation is permitted. Bands: 3.5, 7.0, 14.0, 21.0 and 28 MHz. Messages will consist of Time GMT (must be full 4 figure group) RST and message number (must consist of 3 figure group and start with 001 for 1st contact). Send contest or check log to: Peter Adams, G6LZB, 464 Whippendell Road, Watford, Herts, England WD1 7PT.

APRIL 13-14: The USS Becuna, a World War II Submarine, and the USS Olympia, Admiral Dewey's flagship, will be on the air from 1300Z Saturday to 2000Z Sunday. For a beautiful certificate please send business size SASE to Olympia Radio Amateur Club, PO Box 928, Philadelphia, PA 19105.

THE GUERRI REPORT

Ernie Guerri W6MGI

In the late 1950's the aerospace industry began to implement the idea of using electronics to diagnose and recommend solutions to problems in complex aircraft and missile systems. Because the crew workload in the B58 jet bomber was so demanding, for example, recorded voice warnings were used to alert the pilot to failures in critical systems — engine fires, loss of hydraulic fluid pressure, and so on.

Although this type of non-human intervention is indeed useful, its success depends largely on the premise that design engineers can anticipate all critical failure modes. The fact that many systems experience failures suggests serious flaws in this premise.

A more useful method of fault location and correction would be a scheme that could "learn" the system of which it is a part, and then adapt its actions to respond to the status of the system elements. During the 1970's such methods were implemented using the processing power made possible by ICs and the beginnings of LSI. Termed "expert" systems because they could make rudimentary logical decisions about events that were not necessarily predetermined, such systems proved successful in "fault-tolerant" computers aboard spacecraft in which no repairs would be possible during the planned life of the craft.

In the 1980's, the advent of VLSI has enabled even more dramatic advances: the processing power now available allows the actualization of rudimentary "intelligent" systems. At this point a couple of clarifications are in order. *Intelligent,* in this context, refers to terrestrial carbon-based biological specimens, more specifically, average human intelligence. *Processing power* is a squishy term; Dick Morley, an expert in artificial intelligence, points out that a Cray supercomputer could just about manage the flight dynamics activities of a bumble bee's landing. It takes a lot of processing to handle even the relatively "routine" tasks we take for granted.

Today's intelligent systems generally have the ability to learn from experience — that is, make a statistical determination of the likelihood of subsequent events from a database of prior related events. During the rest of the 1980's we can expect to see increased application of artificial (electronic) intelligence to industrial processes, medical instrumentation, test equipment, and space exploration.

Around the corner — sometime in the 1990's — we will begin to see the first devices with enough processing power to perform actual *inference.* These will be machines with enough memory and sufficient logic to make judgments based on *their own* experience.

The Japanese ICOT supercomputer project is aimed in this general direction, and is expected to achieve its initial goals by the mid-1990's. Similar efforts are underway at several facilities in the United States. Further into the future, perhaps after the year 2000, we can expect to see machines with a greater capacity for inference than humans — probably based on speed and parallel logic — and with non-human logic and communication algorithmic approaches to problem solving. By about 2020 or 2030 we should have machines capable of experiencing emotion — although we may have difficulty recognizing it as such because of its radical divergence from human experience. Perhaps I should add that the machines may have difficulty understanding *our* emotions. Enter bionic shrinks!

The most fruitful applications for such extensive computing power are not yet clear. We must also develop the parallel dexterity which will enable these devices to perform useful work. Each step in the development process will have its advances and setbacks. However, it is nearly certain that we are on the threshold of extending our intelligence to entities capable of transcending our own abilities. We needn't view this prospect as the creation of competitors — or as some might, the creation of our future masters — but rather as an opportunity to expand our horizons.

It's nothing more — and nothing less.

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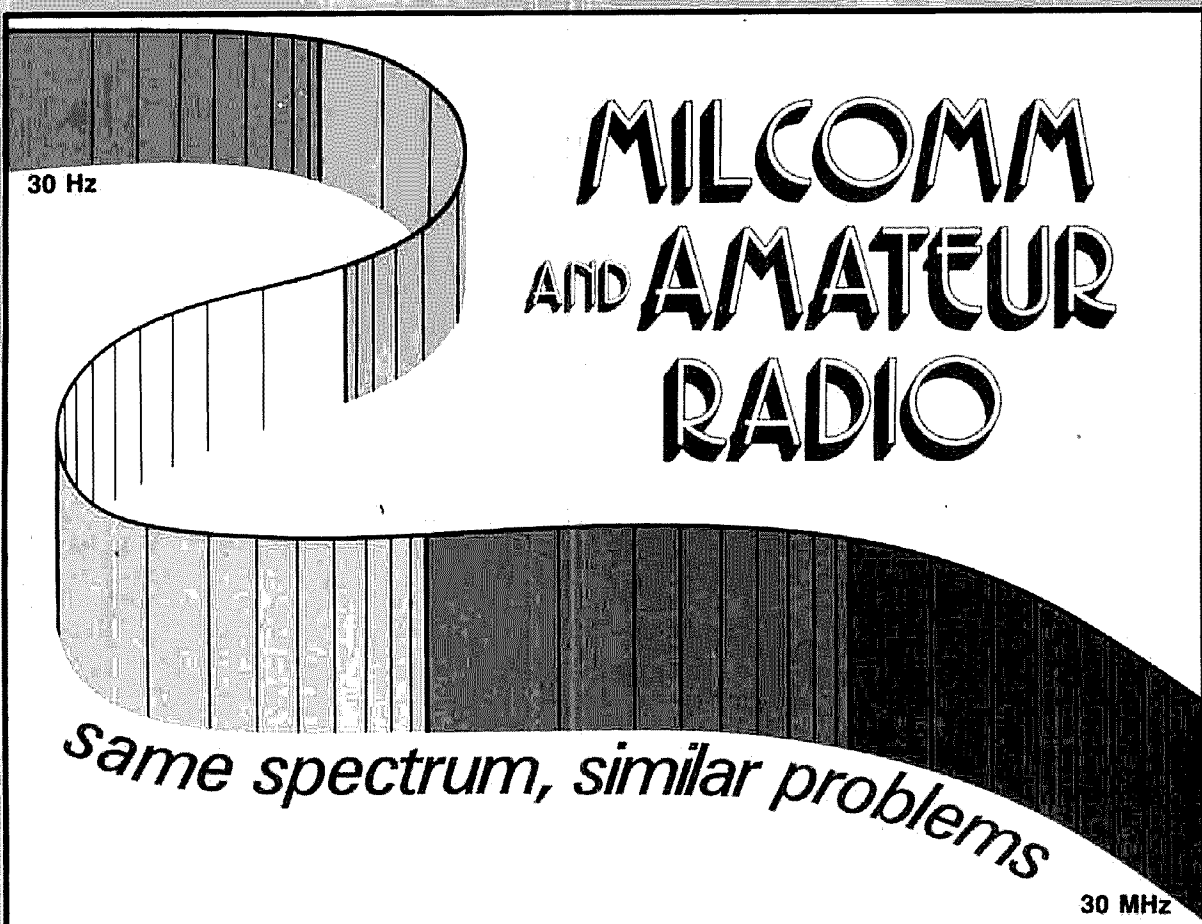

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contents

matrix operation

It started innocently enough. First there was a single response from Europe. "Yes, K2RR, you are being heard here and your signal is 5 and 9." Then someone broke in from the States asking if he could just exchange a signal report with the G station. Before you could say "phased 4-element vertical array" five times, the size of the group on frequency had swollen to 10 Europeans, 1 African, 1 New Zealander, and 400 U.S. stations, give or take a few.

What ensued is what I like to call "matrix operation," whereby a number of stations across the Atlantic get an opportunity to talk to at least *scores* of U.S. stations in very short order. (Consider the phenomenon as a 12 by 400-term matrix and look at the large number of possible combinations and permutations — i.e. QSOs.) This is exactly what happened the other morning on 75 meter sideband during a 2-1/2 hour period in which a number of stations from Western Europe worked rare (for them) states like Nevada, Oregon, and Washington. It also provided many stations in the Western states the opportunity — possibly their first — to talk to Europe on 75 meters.

So what does this prove?

It proves that we **ARE** civilized, contrary to the impression some people might come away with after casually scanning the lower HF bands these days.

These days, more often than not, it appears that some normally decent, law-abiding Amateurs are willing to jam and curse each other out over a piece of precious modern "real estate" — our frequency spectrum — each claiming that he was there first. Others decide to make "critical" adjustments to their transmitters at full power while established QSO's are in progress. The overall result is that what was supposed to be a relaxing hour or so of operating turns into the nearest thing to bedlam, creating ill will — not only among ourselves in the States, but because of the long-distance nature of propagation on the low bands nowadays, other countries as well.

May I be so bold as to share with you some simple suggestions on how to improve low-band operation and keep the collective blood pressure down?

LISTEN *before* transmitting. A local conversation might be taking place on frequency — or a DX contact might be occurring even though you can hear only the closer station.

ASK whether the frequency is in use *after* you listen a bit. It's quite possible that even though you can't hear the stations clearly, they can hear you very well.

ADAPT your operating procedure to the conditions at hand. For example, if a rare DX station is working four stations a minute and you manage to get his attention, don't monopolize him *and* the frequency with a long soliloquy.

ADJUST your equipment for best performance (minimum distortion, spurs, chirp, drift, etc.).

USE the *minimum* power necessary to establish and maintain the contact. Save money: leave your amplifier off as much as possible.

BE CREATIVE. If conditions permit, enhance the rubber-stamp type of DX contact with information of genuine interest — but only if conditions permit (see "ADAPT").

***DON'T* BE CREATIVE.** It is both amusing and sad to hear so many one-way conversations on frequency. The old adage, "If you can't hear them, you can't work them," still holds.

CALL the *least* number of times. Don't try to be the last one heard. Listen to how some top operators do it; sometimes they drop their call in only once — but at the appropriate moment.

DON'T BROADCAST, *communicate*. There are plenty of interesting people out there who have something worthwhile to say.

ENHANCE your knowledge of the band by listening, reading articles on propagation, and noting relationships between the WWV forecasted indices, geomagnetic field status and band conditions.

OBSERVE established "windows." On 75 meters, for example, 3790-3800 is still the international DX window. Your signal at 3800.0001 (LSB), even if perfectly clean, will wipe out the possibility of DX contacts from 3800 down several kilohertz.

LEARN at least a few words in another language. The joy of communicating is yours for the asking.

Though my experience is based on years of operating on 75/80 meters, in general these principles apply to other bands as well. What about your favorite bands and modes? I am very interested in learning about the operating habits, procedures and standard and anomalous propagation modes specific to your band. Drop me a line. Who knows? A cumulative set of notes from these responses could evolve into a pamphlet useful to all.

(Reader's responses to the February editorial "One Million Years of Experience," in which we asked for your suggestions about how the growth of Amateur Radio might be encouraged, continue to pour in. Many thanks to all who've written — a detailed summary of your varied ideas will appear in a forthcoming issue.)

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

AMATEUR RADIO DOES HAVE A FUTURE, BUT A STRONG EFFORT MUST BE MADE to assure that future...that was the consensus of the all-day industry meeting held in Miami January 30, just prior to the Tropical Hamboree. About 40 people, representing many major manufacturers, several distributors, the principal Amateur publishers, and the ARRL attended the session, chaired by former HR Report editor Joe Schroeder, W9JUV.

Amateur Radio Does Have Some Grave Problems, those attending agreed: the static U.S. Amateur population, an uncomfortably high rate of unrenewed licenses, the on-going problems of both manufacturers and dealers, and awareness that our influence in Washington seems to be waning, were just some of the symptoms that were cited. However, during the "off-the-record, no attributions" discussion, a number of really worthwhile suggestions were made.

An Aggressive Program To Attract Junior High School Youngsters was one of the key ideas —junior high science teachers can provide access to this group. Free passes to hamfests for youngsters, their teachers and/or parents was one suggestion, to be complemented by an industry-sponsored and staffed "This Is Ham Radio" booth to provide both an introduction and even "hands-on" experience for newcomers. Supplementing this effort will be an Amateur Radio "comic book" highlighting its "fun" aspects in an entertaining way. Also planned are "sales pitches" to the general public, directing a special effort toward responsible CB groups such as REACT and scanner organizations. The possibility of making the entry level more attractive, by adding limited Novice data and/or voice privileges, was also discussed. Greater dealer involvement, such as hosting training courses and club meetings, was also considered. A proposal that industry representatives sit on the ARRL board, to encourage closer ARRL-industry coordination, was broached to ARRL representatives.

Overall Attitude After The Grueling Session Was Upbeat, with those attending feeling that some very real progress had been made in overcoming Amateur Radio's current inertia. Volunteers from the group are already actively working on a number of the suggested programs, and progress reports on their efforts plus a discussion of future plans is scheduled for a Dayton meeting the Thursday evening before the Hamvention.

FCC'S REPEATER FREQUENCY COORDINATION PROPOSAL will place prime responsibility for resolving repeater conflicts squarely on operators of uncoordinated repeaters. In the case of a dispute with two coordinated machines, both operators would share responsibility for its resolution equally. In its Notice of Proposed Rule Making, the FCC cited the rapidly increasing number of repeater interference complaints it's received, and that the bulk of these problems seem to involve uncoordinated repeaters.

The FCC Asks Amateurs To Consider A Number Of Key Questions in this potentially far-reaching NPRM. For example, should coordination be mandated in major urban areas? As an alternative, should narrow-band technologies and tone squelch be required to minimize the interference problem? Should the Commission recognize a "single national frequency coordinator," either on a national basis or as an "advisor" to local coordinators?

A Blanket Moratorium On New Repeaters Initially Distracted Attention from the real issues until lifted by the FCC February 19. There had been some comment that the moratorium had simply confused the issue and was probably unenforceable anyway, and the ARRL (supported by the Tri-State Repeater Council) had petitioned the FCC to rescind the ban.

Comments On The Repeater Coordination Docket, PR Docket 85-22, are due at the FCC July 1; Reply Comments will be due September 30.

TEXAS HAS JOINED THE SHIFT TO 20 KHZ 2-METER SPACING, adopting the change by an 8:1 margin at the February 16 meeting of the Texas VHF FM Society. The decision, which makes Texas the second state east of the Rockies to make the move, has been under consideration for some time. As yet, no timetable for the move has been established.

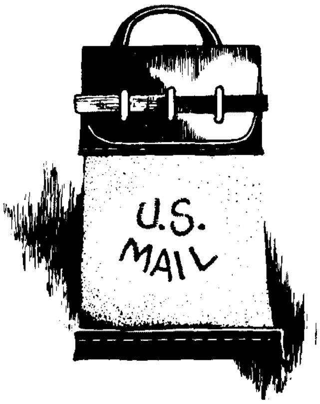
Northern California And Minnesota Are Also Looking At The 20 kHz Plan. In California the Northern Amateur Relay Council has invited Clay Freinwald, K7CR, to its April 13 meeting at Concord to discuss the 20 kHz plan. K7CR is considered one of the fathers of the plan in the Pacific Northwest. Minnesota has 20 kHz on its April meeting agenda, too.

The Escalating Move To 20 kHz Is Causing Concern In Europe where 2-meter FM operation on their 144-146 MHz band has traditionally been on 25 kHz centers. Some Europeans are worried that the Japanese, whose domestic 2-meter band is also on 20 kHz spacing, would drop 5-kHz step capability from their rigs if the U.S. market no longer needs it.

TWO ESTABLISHED REGIONAL VECS ARE GOING NATIONAL. Both the DeVry Amateur Radio Society and Metroplex have applied for national status, based on their experience and success in operating regional programs in the Ninth and Second call areas, respectively. The DeVry proposal is particularly interesting, as DeVry has campuses in seven other call areas that could serve as nucleii for its efforts in those areas. As the FCC's experience with both organizations has been excellent, their proposals should be accepted quickly.

WARC BAND EXPANSION IS STILL HANGING FIRE, with some GMRS representatives looking into the possibility of carving a slice for personal radio from the new 902-928 MHz band.

The Commission's Recent Turndown Of A Personal Radio Service at 900 MHz has also been challenged by the same group, who've filed a Petition for Reconsideration with the FCC.



comments

Amateur Radio — not what it used to be?

Dear HR:

As a lifetime subscriber, I have a few comments about K2RR's account of his maritime experiences ("Reflections," November, 1984, page 5). I was in the Merchant Marine during World War II, but as a fireman/water tender/oiler. I was discharged the day I was to get my commercial ticket and never went to sea again.

K2RR only hinted at what is missing from Amateur Radio today. Gone is the clean language, the exchange of technical information, the invitations to visit (especially to mobiles), the hospitality we used to have, and in short, good manners on the air.

I'm glad I was born sooner and had the opportunity to enjoy Amateur Radio for the last 48 years. I had it at its best.

Albert Kaufman, W1JVQ
Bridgeport, Connecticut

grid dipping

Dear HR:

Even though George A. Wilson, Jr., W1OLP, in "Matching Dipole Antennas," (May, 1984, page 129) made at least 24 separate references to GDO (Grid Dip Oscillators) and Grid Dipping, someone is certain to try substituting a solid-state dipper, (such as the Heathkit HD-1250 or one of several factory assembled versions) when exciting the RF Bridge discussed in the article. In fact, with the solid-state dipper far more prevalent today than the old vacuum tube grid dip oscillator (and interchangeable in most applications), no doubt a large number of hams who build the RF bridge will end up frustrated and with no discernible "dip."

While the solid-state dippers can be

user group agrees on 23-cm band plan

Over 100 users of the 23-cm band (1240-1300 MHz) reached agreement on an updated regional plan to serve the needs of the southern California area for the next three years at a meeting in Orange County, California. Present and participating at the meeting, sponsored by the Southern California Repeater and Remote Base Association (SCRRBA), were representatives of all users' modes currently operational, or likely in the future to require spectrum, on 23 cm.

Each of the 23-cm band users' groups (ATV simplex, ATV repeaters, weak signal/experimental, FM voice repeaters/links, digital, satellite/AMSAT, VRAC, VUAC, and SCRRBA) selected a representative to participate in a four-hour roundtable and negotiation session. The resulting plan — basically a modification of the existing plan — was prepared specifically for use in their region. Key stipulations provide that the plan will remain in effect for the next three years, after which time a similar meeting will again be held to review the existing band utilization patterns, and that the frequency allocations of existing users' groups will change only as new users' groups begin to operate on the band.

Because the region leads the nation in terms of 23-cm band activity, SCRRBA suggests that the band plan shown in **table 1** may be useful to other coordination councils in preparation of their own regional band plans.

— SCRRBA

table 1. A comparison of present and modified 23-cm band utilization plans (courtesy SCRRBA, P.O. Box 5967, Pasadena, California 91107).

band segment	present usage	new usage (initiated as needed)
1240 - 1246	ATV repeater (Channel 1) (1241.25 video carrier, VSB filtering required)	ATV repeat (Channel 1) (1241.25 video carrier, VSB filtering required)
1246 - 1248	narrow-band FM point-to-point links (voice)	same plus narrow-band digital (<50 kHz BW)
1248 - 1258	ATV repeater (Channel 2) video carrier 1253.00	
1248 - 1251.5		wideband digital (>500 kHz BW)
1251.5 - 1252		guard band
1252 - 1258		ATV repeater (Channel 2) video carrier 1253.25, VSB filtering required
1258 - 1260	narrow-band FM point-to point links (voice)	same plus narrow-band digital (< 50 kHz BW)
1260 - 1270	satellite uplink ATV repeater (Channel 4) 1277.00 video carrier	satellite uplink, plus non-coordinated simplex: *experimental* wideband, no repeater inputs/outputs
1270 - 1272	FM (voice) repeater inputs; 1271.000 "test pair" input	FM (voice) repeater inputs; 1271.000 "test pair" input
1272 - 1282	ATV repeater (Channel 4) 1177.00 video carrier	
1272 - 1275.5		FM repeater future expansion, ACSB systems, linear translators

used to determine resonance, per the first part of George's article, it is not likely to provide enough excitation to obtain a reading with the RF bridge unless overcoupled, with sensitivity set at maximum, and with an extremely sensitive μA meter used as the detector. Even a 50 μA meter will probably not allow a discernible "dip" to be obtained!

A rough idea of a dipper's suitability can be obtained by connecting a germanium diode and a small 2 to 3-turn link in series across the μA meter's terminals. Coupling the link to the dipper's coil should easily produce a full-scale reading. If it does not, the dipper cannot be used to excite the RF bridge.

Robert G. Wheaton, W5XW
San Antonio, Texas

receiver input temperature

Dear HR:

Amateur Radio literature never mentions (or I've never seen) the temperature contributed by a transmission line to receiver input temperature. None of this mattered with hot, noisy amplifiers, but it could matter significantly with the low-temperature amplifiers coming into use. If someone can come up with a convincing argument that the transmission line adds no temperature of its own, I'd like to see it. Such a line would make an ideal cold-source for noise measurements.

Picture the line as a string of small attenuators generating their own thermal noise, which is sent in both directions along the line. The noise energy has an equivalent temperature at the line's Z_0, which is some fraction of the line's ambient temperature.

I derived the temperature output for a small attenuator (0.1 dB, 6.769 degrees K) and summed the cumulative temp (with cumulative losses), generating a chart of temperatures for various losses. It closely follows the equation

$$T = 298 \left[1 - \text{antilog} \left(\frac{-dB}{10} \right) \right] \text{degrees}$$

I deliberately ignored the possibility of any energy reflected from the *antenna*.

This sugggests that 30 feet of RG-59B/U could add 127 degrees K at the input to a 432-MHz amplifier, and 6 feet would do the same at 1296 MHz (approximately). 15 feet of 1/2 inch hard-line could add 36 degrees to the input at 1296 MHz.

I'd like to hear if anybody has any *thoughts* or can suggest any references on this.

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ham radio

digital HF radio:
a sampling of techniques

Tomorrow's technology is at work in the military today

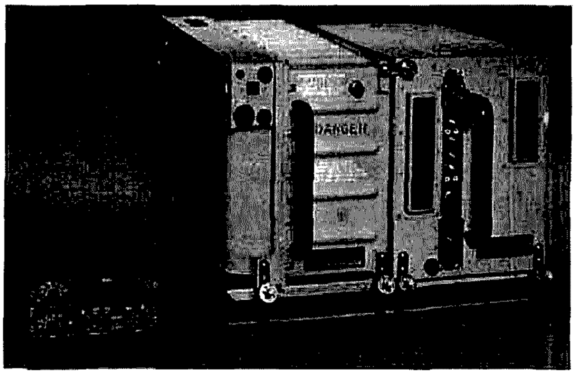


fig. 1. P-3C AN/ARC-161 HF radio set.

Thinking of military communications equipment, one conjures up the image of sturdy, olive drab-colored, compact all-inclusive designs that meet stringent operating requirements. On the other hand, when one thinks of Amateur Radio equipment, a different picture comes to mind. Instead of the predictable "military" designs, one imagines a diversity of commercial homebrew items with little uniformity of design or form.

But which spectrum environment — military or Amateur — is more congested? Which demands more stringent operating requirements? The answer is not obvious.

From his unique vantage point as president of Communications Consulting Corporation, Dr. Rohde, an experienced receiver designer and active ham provides insight into how military equipment designers solve their problems of congested spectrum using the most sophisticated techniques and materials. Some circuits will doubtless appear in Amateur Radio applications — perhaps a few may already have done so.

— Editor.

Until recently, HF radios for the military (fig. 1) have been designed and built in the traditional analog way, with selectivity obtained through the use of LC or crystal filters in the IF section and active filters in the audio frequency section. These radios have been used for point-to-point operation where only infrequent change of operating frequency was required. In addition, these point-to-point connections were used with constant output power.

In order to meet communications goals for 1985 and beyond, particularly in military applications, modern HF equipment must be adaptive, frequency agile and capable of supporting secure digital voice communications. It must be capable of operation on both a point-to-point and networked basis as well. Adaptivity is needed both to control transmitter power to a level no greater than required for the connection, and to select frequencies which provide good propagation with a minimum of interference.

Frequency agility is desirable in the event of deliberate jamming or rapid change of propagation conditions. Transceivers must have the ability to change *frequency* rapidly enough to adjust to changes in environment. While propagation conditions normally change relatively slowly, avoiding a jammer requires a repetitive fast change of frequency called *frequency hopping*. For a jammer to be effective in disrupting communications, it must either be extremely powerful, use a high-gain antenna, and cover a wide frequency range, or operate narrowband and try to predict or detect the frequencies on which the frequency hopper is operating and jam only those.

By Dr. Ulrich L. Rhode, KA2WEU/DJ2LR, 52 Hillcrest Drive, Upper Saddle River, New Jersey 07458

Generally, the optimum hop rate for communication is determined by a trade-off between implementation cost and a combination of operating considerations which include expected propagation delays (related to distances between jammer and communicators), capabilities of enemy direction-finding and jamming equipment, and required communications distances, bandwidths, and reliabilities. Taking all of the above into account, recent Army, Air Force, and Navy requirements have focused on hop rates in the range of several hundred to several thousand hops per second as needed to satisfy most military situations (actual rates are classified information). Frequency-hopped signals faster than 1 millisecond or so are difficult to locate with currently-deployed tactical direction-finding hardware and would require a costly "Fast-Follow" jammer to track and jam the communications on a hop-by-hop basis.

other processing techniques required

Often frequency hopping alone is insufficient to defeat deliberate jamming. Modern HF transceivers must be equipped with sophisticated signal processing techniques that provide several levels of redundancy and/or error correction capabilities. These techniques, involving more complex circuitry and larger instantaneous bandwidths, allow recovery of desired signals even in the presence of high levels of natural noise or deliberate jamming. The modern HF transceiver has an RF portion which must be "transparent" to the real brains of the transceiver, the digital signal processing circuits.

Finally, military HF radios must be designed to support the transmission of secure voice and data. Although many techniques exist for manipulating analog voice signals to provide privacy, the U.S. military has settled on the encryption of digitized voice as being both easier to accomplish and more secure. Also, to overcome frequency selective fading characteristics of HF, techniques for digitizing voice have included bandwidth compression as well. This is because narrowband signals suffer less distortion. What is sought is the lowest possible bit rate that produces acceptable speech quality when converted back to its analog form.

voice encoding technique

Linear predictive coding (LPC is the compression scheme currently favored by the U.S. military for HF links. Used at a bit rate of 2.4 kbps in the ANDVT (U.S. Navy Advanced Narrowband Digital Voice Terminal), LPC encodes the voice as numbers derived from the instantaneous spectral characteristics of the voice. The numbers themselves have no relationship in an analog sense to the original voice signal, but they are used in an inverse process to produce an approximate analog signal resembling the original voice. The process is not unlike that used in children's toys that speak (e.g., "Speak and Spell"™).

Present military standards have settled on LPC-10, a linear predictive coding/decoding algorithm which compresses 3 kHz speech to a 2 kilobits per second data stream. The algorithm uses a linear mathematical relationship to predict the value of each successive sample it is digitizing and hence the name.

To make the compressed data secure, it is encrypted by combining it in a unique mathematical fashion with a string of numbers generated by a "key" generator. The result is a new succession of data carrying voice information encrypted by the "key" and the method of combining it with the voice data.

At the receiver, demodulation depends on synchronously detecting the transmitted bits even though they cannot yet be converted to intelligible voice. Then, by proper mathematical application of the same "key" used to encrypt the original data, the decrypted bit stream is applied to the inverse of the LPC process, and the spectrum-related numbers are converted back to analog speech.

The entire process depends heavily on maintaining good channel quality so that accurate bit timing at the receiver may be achieved; the most important aspect is to preserve timing relationships. A decision at the receiver as to whether a received symbol (character) is a "one" or a "zero" must not be made during a symbol transition period. As the data rate increases it becomes more difficult to insure that the receiver is making its decisions at the proper time. Decision errors, the result of distortion that occurs between characters (inter-symbol distortion) cause degradation in the overall system performance. Timing relationship distortions are the result of frequency selective phase shifts that arise from differences in path length as the signal component at each frequency passes through and reflects from different layers of the ionosphere. The narrower the instantaneous bandwidth of the radiated signal, the fewer the perturbations. Consequently, the best results are achieved by using the narrowest instantaneous bandwidth design.

With simple forms of modulation such as a Binary FSK, Bi-Phase FSK, or even AM, 6 kHz or more bandwidth would be needed to send the 2.4 kbps LPC signal. To fit the signal into a standard *3 kHz* voice channel requires more sophisticated modulation schemes. For example, to achieve 1 Hz/bit packing density, ANDVT uses a 39-tone parallel modulation scheme. The 2.4 kbit data stream is split into a number of parallel data streams transmitted at a lower rate. According to the ANDVT algorithm, each of the slower streams modulates one or more of the 39 tones.

The resulting tones are summed in an analog summer and applied as conventional analog modulation at the exciter.

This parallel modulation approach trades complexity and transmitter efficiency for narrow bandwidth. The transmitter (at maximum power) has to be able to handle the case in which all tones add in-phase (maximum power out) and, as a result, must operate well below peak output power most other times. Complexity results from having to handle parallel channels both at the transmitter and receiver. At the receiver, careful tracking of frequency and phase is needed to insure that demodulation of each tone occurs properly. Further, for the frequency hopping case, differences in path length at each frequency must be compensated for on a hop-by-hop basis. Nevertheless, these techniques do work and are being incorporated, primarily in software-based modems into new military HF equipment.

An alternate approach under study at RCA is a serial modulation scheme which processes blocks of data taken in sequence from the data stream. Each block is encoded as one of several tones. If 6 bits of data are taken at once and the modem has 64 (2^6) available tones in a 3 kHz audio bandwidth to pick from, then each 6 bits of data determines which one of those tones is to be transmitted. The selected tone is transmitted with a duration of roughly six times the original bit duration; the transmitter operates at full output power; and the instantaneous bandwidth, once the system is synchronized, is one-sixth that of the parallel scheme. This approach has been implemented and tested at RCA. Its penalties include a complex synchronization and demodulation algorithm and a rather unique (and objectionable) on-air "signature."

key digital building blocks

Traditional analog circuit techniques are inadequate to meet the previously stated requirements. The modern HF transceiver must be based not only upon new intelligence, but on a new set of "building blocks" as well. Many of these building blocks are digital functions performed entirely by software routines; others, involving high computational rates, are better implemented by using dedicated digital hardware. Some of the more obvious building blocks include the following:

Modems. Modems, in general, are required to process voice or data and to provide the necessary waveforms, usually at IF, to the exciter and receive them from the receiver. Selection of the waveform is critical since performance of the entire link depends on the features it possesses. The waveform is said to be "robust" if it has at least two levels of redundancy in its synchronization scheme. It must also have built-in error detection and correction as well as preambles, which allow channel quality measurements. In addition, the waveform may have to support a variety of data rates to maintain high quality information transmission as link conditions degrade.

Digital filters. For reasons of flexibility and performance, digital filtering is preferred in the IF and audio frequency sections of the transceiver. The ability to choose an appropriate filter type and time constant greatly simplifies modem design. Also, under software control, it is possible to have a digital filter assume any of the classic shapes or performance characteristics as required. Although either Chebyshev or Bessel characteristics are generally used, others can be selected. On the negative side, use of sophisticated digital filters sometimes requires correction of group delay effects of any remaining LC or crystal filters in the transceiver. Also, real time digital filters require correspondingly fast high-resolution A-to-D and D-to-A converters as well as high throughput microcomputers. Two devices ideally suited to digital filter implementation are the Texas Instrument TMS-320 and the RCA (internally developed) high performance ATMAC II CMOS/SOS Processor.

Frequency synthesizers. Sufficiently fast-switching frequency synthesizers cannot be easily implemented in the traditional analog form using multi-loop PLL synthesizers; instead, the direct digital synthesizer (DDS), to be described later, is required. The DDS features arbitrarily fine resolution and uses a cosine look-up table together with a microprocessor, a D/A converter, and a lowpass filter to generate such waveforms. The DDS can be built with switching times between 1 μs and 50 μs, with the actual switching time dependent on the D/A converter and the number of glitches produced by the sampling and integration process. A limiting signal-to-noise ratio of about 75 dB, which has to do with the sample rate, is the current state-of-the-art.

The direct digital frequency synthesizer output is mixed with the output of a conventional single-loop PLL synthesizer with wide loop bandwidth. This PLL then determines the settling time of the overall system.

Agile antenna couplers. Since wideband antennas aren't commonly used, the transmitter or receiver has to be matched to the antenna on a hop-by-hop basis. The switching speed of the antenna coupler is thus part of the overall system switching time. Conventional relay couplers can be built with 10 ms switching speed relays, and there are claims for future reed relays that will provide 1 ms switching times. However, if 10-100 (or more) frequency changes or hops per seconds are required, the lifetime of the mechanical devices will be soon depleted. The typical lifetime of these devices is about 2 million operations or, at 100

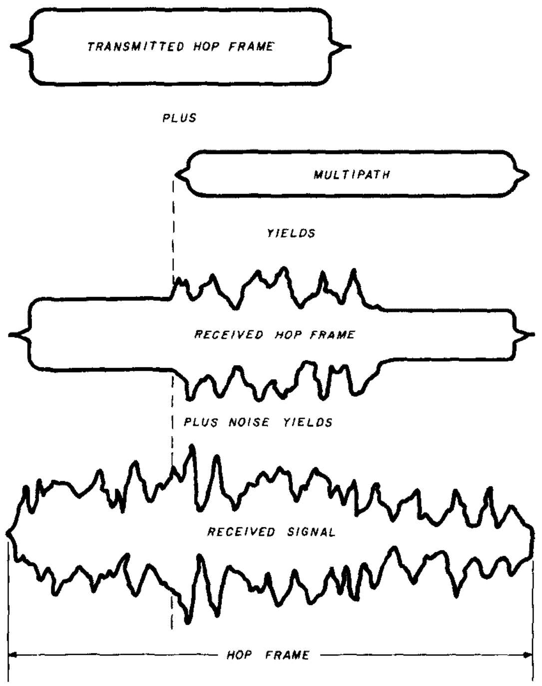
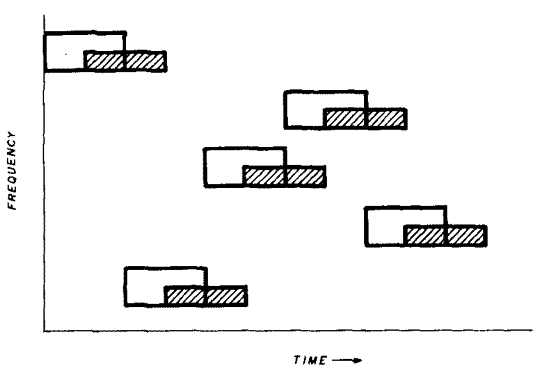


fig. 2. Frequency hopping multipath problem.



Note: Illustration assumes multipath delay equals one-half hop frame. Amplitude is one-half of desired signal.

fig. 2(A). Frequency hopping with multipath.

switches per second, 20,000 seconds — only three hours of continuous use.

Modern couplers, then, must employ solid-state switching, for which the use of PIN diodes is the obvious choice.* Such a coupler uses a quasi-binary coded arrangement of inductors and capacitors that are switched in and out of the matching network by PIN diodes. Because antenna couplers have to be built to handle power levels of up to 1 kW, as much as 8,000 volts DC is required for reverse biasing; forward currents of up to 2 amps are necessary. In order to generate these voltages and currents, special dedicated switching power supplies have to be provided, and a mechanism is needed to bring the voltage or current to the switching diodes without introducing parasitic stray effects.

Having summarized both the basic requirements and some of the key building blocks for modern digital HF radios, we'll now examine some specific design approaches.

frequency hopping receiver design

The need to frequency-hop at HF, especially in the presence of multipath, places the most stringent requirements on the modern HF transceiver. The following section discusses frequency hopping, design considerations, and introduces the unique problems created by frequency hopping. Several means of solving these problems are considered, and a novel solution illustrating the use of digital techniques is presented.

Frequency hopping in a multipath environment. The effect of multipath distortion is depicted in **fig. 2.** A transmitted hop frame represents a burst of signal energy radiated at one frequency and received by a frequency hopping receiver (synchronized to the transmitter). A *multipath* is a burst of identical signal energy, delayed in time and reduced in amplitude. The number of multipaths, amount of relative time delay, and attenuation of the received burst depend on link characteristics such as frequency, separation, and time of day.

What the receiver actually "sees" during its dwell time (window) is a combination of all the energy contained in all the multipath bursts on the frequency. This is illustrated in **fig. 2,** both with and without noise. This burst distortion will be different for each frequency while frequency hopping. The receiver must be able to correctly demodulate the transmitted information in this distorted signal.

Intersymbol interference reduction techniques. If a transmitter is sending a succession of symbols (such as ones and zeros), the receiver's job is to identify those symbols and convert them back to intelligible information. When a signal includes additive noise, errors occur because the difference between the two levels is not as clearly defined. When multipath corruption occurs, symbols actually overlap and intersymbol distortion occurs. This is a special type of distortion which requires more sophisticated processing to overcome. Inter-symbol distortion caused by multipath is generally what limits the throughput capability (maximum data rates) of HF links. A number of techniques have been tried to reduce inter-symbol in-

*See "High Power RF Switching with PIN Diodes," by J.R. Sheller, KN8Z, *ham radio*, January, 1985, page 82.

table 1. Current HF signal processing techniques.

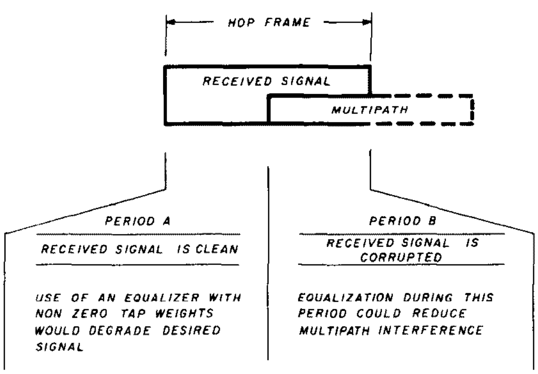
technique	concept	remarks
parallel tone differential quadriphase shift keying (DQPSK) modulation	long bauds on adjacent frequencies — nonadaptive	selective fading causes high error rates high peak-to-RMS ratio transmitter
parallel matched filters or rake processing	pulse matched filter (correlation)	good for low data rates or when WT > 30
linear equalization	minimize distortion by filtering the received signal	severe multipath causes high error rates requires training sequence and updating
decision feedback equalization	minimize distortion by filtering the received signal and past decisions	tracking problem, requires training sequence and continuous update
maximize likelihood estimation	message matched filter viterbi decoding algorithm	exponential growth with multipath delay tracking problem

terference, with the most common method shown in **table 1**. We will concentrate here on equalization.

Equalization techniques. Figure 2A illustrates a simplified model of frequency hopping with multipath. It shows a frequency-versus-time plot of a pseudorandomly hopped signal. Each frequency is represented by a received signal component and a cross-hatched multipath signal of one-half amplitude and delayed half the hop frame. Let's look at the problem of implementing an effective equalizer in this hopping environment.

When a single frequency hop frame is examined (assuming receiver synchronization), as in **fig. 3**, the received signal is present for the entire hop frame, while the multipath signal may be observed only during part of the hop frame, identified as period *B*. During period *A*, the first part of the hop frame, the received signal is not corrupted by multipath components, and one would expect only normal atmospheric and receiver noise to be present. Trying to eliminate multipath during this period using feedback equalization would actually degrade the desired signal. In fact, *no* special processing should take place during period *A*. During period *B*, however, the signal is corrupted by multipath and some processing may be used to reduce the multipath interference.

But there is a further complication in that a single communications link or net is not necessarily the sole user of a frequency band or a family of hopped frequencies. This is especially true in the crowded HF band. Consequently, the simplified model must be extended to include a large number of users who may share the same frequencies. The shared use of frequencies by several synchronized nets is illustrated in **fig. 4**.



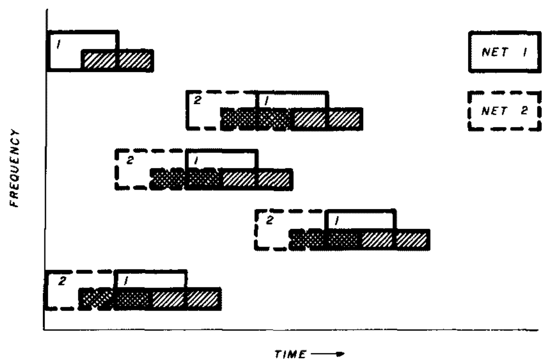
NOTE: ASSUME SIGNAL WITH SINGLE MULTIPATH INTERFERENCE COMPONENT.

fig. 3. Simplified model.

Additional "friendly" stations cause problems. Multiple users in a network present a more complicated interference pattern at the receiver. This is illustrated in **fig. 5**. The hop frame contains the desired signal and two distinct and different interference components. During period *A* there is a new delayed replica of an undesired signal from another user or from another net. This signal has totally random parameters with respect to the desired signal.

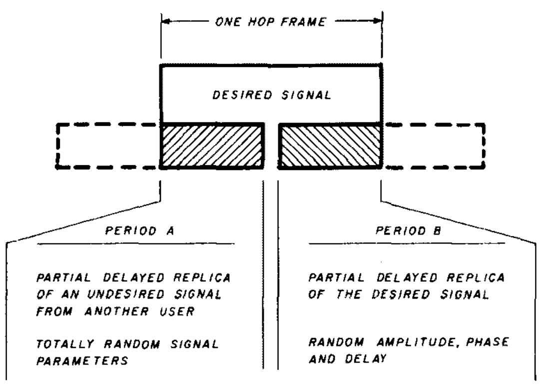
Also if the frequency hopping rate is increased to "outrun" the multipath, interference will be present from the multipath components of other synchronous net users and from others sharing the frequency.

In **fig. 5**, each hop frame is shown as consisting of the desired signal and several different types of interference. During the first part of the hop frame, (period A), random interference is present which does not cor-



Note: Illustration assumes multipath delay equals one-half hop frame. Amplitude is one-half of desired signal.

fig. 4. Two-net frequency hopping with multipath.



Periods *A* and *B* may overlap or have a gap since multipath for each user is unique.

fig. 5. Frequency-hopping interference.

relate with the desired signal. The only known technique to reject this type of interference, which may have the same waveform characteristics as the desired signal, is to use correlation processing, which rejects uncorrelated random noise. This type is processing, achieved by deliberately spreading the spectrum of the original signal at the transmitter and compressing it again at the receiver, cannot be easily achieved at 2.4 kbps data rates. This is because the usable channel bandwidth is already fully occupied by the non-redundant digitized and compressed voice signal. To maintain occupied bandwidths at 3 kHz or so while transmitting data at 2.4 kbps, time-gated adaptive equalization may be used. The name time-gated is applied because it should only be active during period *B*, when the multi-path burst *is* a replica of the desired signal. During period *B*, even a high speed dedicated processor will be taxed. For each multipath component, the individual tap locations must be determined. When frequency hopping over a 10 percent bandwidth, tap location variations corresponding to relative multipath delays may be small, but the amplitude and phase weights differ for each frequency hop frame. Weights may also have to be varied from hop to hop due to doppler shifts encountered on the channel — the result of ionospheric variation, as well as motion of the user. Because of these effects, amplitude and phase of the tap weights cannot be determined only once and revised each time the same frequency is revisited. They must be determined for *each* hop. Prior data may be helpful as an initial estimate, but cannot be relied on for adequate equalization.

sample design approach to multipath processing

An equalizer is actually a matched filter that attempts to model the corrupted channel as a function of time. The input to the equalizer is a composite waveform consisting of wanted signal, noise plus distortion components. The approach to be described is based on an RCA-developed adaptive algorithm which repeatedly calculates the ratio of desired signal to signal plus noise plus distortion. Maximizing the ratio by rapidly adjusting characteristics of the equalizer comprises the adaptive process. When the ratio is maximized, the equalizer is said to be "converged to the value best representing the inverse of the corrupted channel." Passing the corrupted input signal through the adjusted equalizer essentially removes the corruption and gives the best possible signal. The process is very much like that used to equalize trans-Atlantic telephone cables, except that in this process the equalizer characteristics must be adjusted each time the transmitter-receiver pair hops to a new frequency.

The basic operation consists of storing all signal samples received during the dwell on a particular frequency. The digital processor uses the samples to compute the equalization measure described above until a final value is found. In reality, the equalizer resembles a tapped delay line with complex weights applied to the signal developed at each tap. The taps are summed and, using an appropriate algorithm, added to the original input signal. An equalization value is computed for each set of tap weights examined. The tap weights producing the highest equalization value are then applied to the equalizer and the original set of signal samples filtered and passed on for further processing.

For binary FSK signals, the equalization value can be found from spectral energy measurements using a Fast Fourier Transform (FFT). The process consists of taking the spectral energy in the two carrier frequencies and comparing it to the energy in the remaining in-band spectral region after the carrier frequencies have been deleted. The equalizer is optimally adjusted

when the desired carrier frequency energy, less residual energy, is maximized. Thus, the adaptive process would vary the complex tap weights until that difference is maximized.

Appropriate techniques can be found for other forms of modulation and will be discussed subsequently, but in general, they all compare signal energy in a known region of the spectrum (where the desired signal should have the majority of its energy) to the remainder of the spectrum where multipath components lie.

The RCA approach is unique in that it does not require a known signal to be transmitted at the outset of each new hop frame. Such a signal (termed a "training" signal) reduces the usable throughput of the system since it occupies a portion of every hop frame.

A second attractive feature of this equalizer approach is that any arbitrary signal possessing the desired waveform characteristics (i.e. modulation type and rate) may be automatically equalized by applying the same algorithm. This allows a receiving station in a network to automatically equalize signals from any other station in the network without first having to identify which station it is or decode a special "training" signal.

A block diagram of the multipath processor is illustrated in **fig. 6**. Input signal samples provided by the frequency hopping receiver are stored in an input buffer. When an entire hop frame has been accumulated, the samples are transferred to a second buffer, which stores all the signal samples associated with two hop frames. While the input buffer is collecting new data from a different carrier frequency at the next hop, the *output or hop frame buffer* is used repetitively to perform the equalization. For a frequency hop rate in the low hundreds, there is sufficient time to recirculate the stored data through the equalizer and adjust the appropriate tap weights for highest channel quality.

The output hop frame buffer supplies digitized signal samples to two separate processing functions — the time-gated equalizer and the tap locator. The time-gated equalizer is controlled by the tap locator and the tap weight calculator. After an "equalization complete" signal is provided by the tap weight calculator, the signal samples stored in the receiver signal buffer are passed through the now adjusted equalizer and the corrected data is stored in the output buffer. The signal samples are supplied on demand to the next processing stage, which may be a spread spectrum despreader or data demodulator.

Time-gated feedback equalizer. A time-gated feedback equalizer is shown in **fig. 7**. Its structure is very similar to that of any other feedback equalizer. It has an input signal, a differencing circuit, and a weighting network. The difference is supplied as an output signal to the data demodulator. The delay line is a multilevel shift register equal in length to the hop frame. The tap spacings, corresponding to minimum resolvable multipath delay, are equal to the signal sampling interval, which is considerably smaller than one bit in duration.

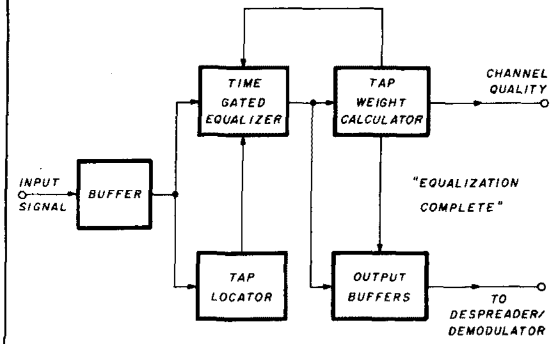
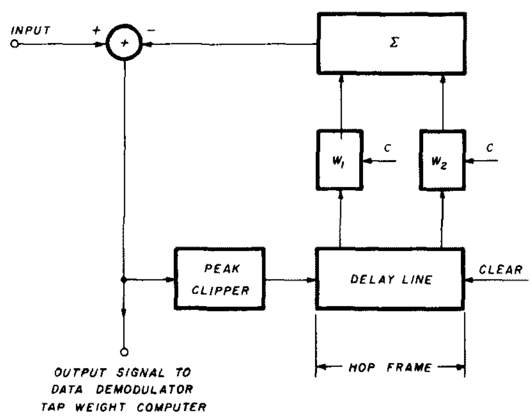


fig. 6. Multipath processor.



NOTES:
1. CLEAR DELAY LINE AT BEGINNING OF EACH HOP FRAME (STORE ALL ZEROS).
2. CLOCK OUTPUT SIGNAL INTO DELAY LINE.
3. WHEN OUTPUT SIGNAL REACHES FIRST DELAY LINE TAP, WEIGHTS W_1 (AMPLITUDE AND PHASE) BECOME EFFECTIVE.
4. SUBTRACT MULTIPATH COMPONENT FROM INPUT SIGNAL.
5. REPEAT FOR SECOND MULTIPATH COMPONENT USING WEIGHTS W_2.

fig. 7. Time-gate feedback equalizer.

The delay line in **fig. 7** is tapped in two places. Two multiplying weights, W1 and W2 are shown; these are complex and represent both amplitude and phase weights. The resulting outputs are summed and fed back to the input differencing circuit. The peak clipper is incorporated to prevent positive feedback for certain data sets at high multipath levels.

Time-gating the feedback equalizer is achieved by clearing or resetting the delay line to zero at the beginning of each hop frame. Consequently, when a new hop frame starts, there is no feedback. In fact, there will be no feedback until the first samples reach the first tap. At that point, feedback of the appropriate amplitude and phase will begin cancelling the first

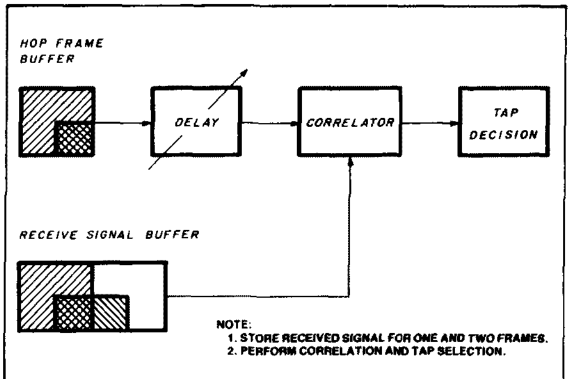


fig. 8. Correlation process.

multipath component. (A second multipath component will not become bothersome until the output signal reaches the second tap with amplitude and phase weight W2. Up to that time the finite values of amplitude and phase weights of weight W2 will have been multiplied by the zeros advancing in the delay line and the tap will not have been effective.) After that time, feedback cancels the second component, and so on. The impact of a large number of taps is that additional time is required to optimize the amplitude and phase weights. It is reasonably assumed from HF path predictions that only three to five multipath components will significantly affect a signal. The hardware necessary to process this data is available right now.

Tap locator. The second function in the multipath processor is the tap locator. The receiver provides two signals for use in an auto-correlation function which determines multipath tap locations. Once tap locations are known, the tap weights for the filtering function can be adjusted. This correlation is illustrated in **figs. 8** and **9**. The original transmitted frame is transferred to two buffers. The contents of the two buffers are then fed to a correlator.

The output of the correlator is examined for correlation peaks above a predetermined threshold, which indicates that a multipath component is present at that particular value of delay. This auto-correlation process determines the multipath delay without depending on any external timing. It is not affected by timing uncertainty or by jitter from one frequency hop to another.

Tap weight calculator. The third major function in the multipath processor is the tap weight calculator, which consists of a channel quality measurement unit and a tap weight programmer. The tap weight programmer alters the amplitude and phase of the tap weights in response to instantaneous channel quality measurement. Because hop frames are buffered, the quality of the channel can be measured repetitively as the equalizer converges and used to control the tap weights in a closed loop fashion. The quality measurement is based on an RCA-patented quality monitoring technique particularly applicable to a variety of angle modulated signals. It does not require any special training signals, but instead operates directly on the data signal.

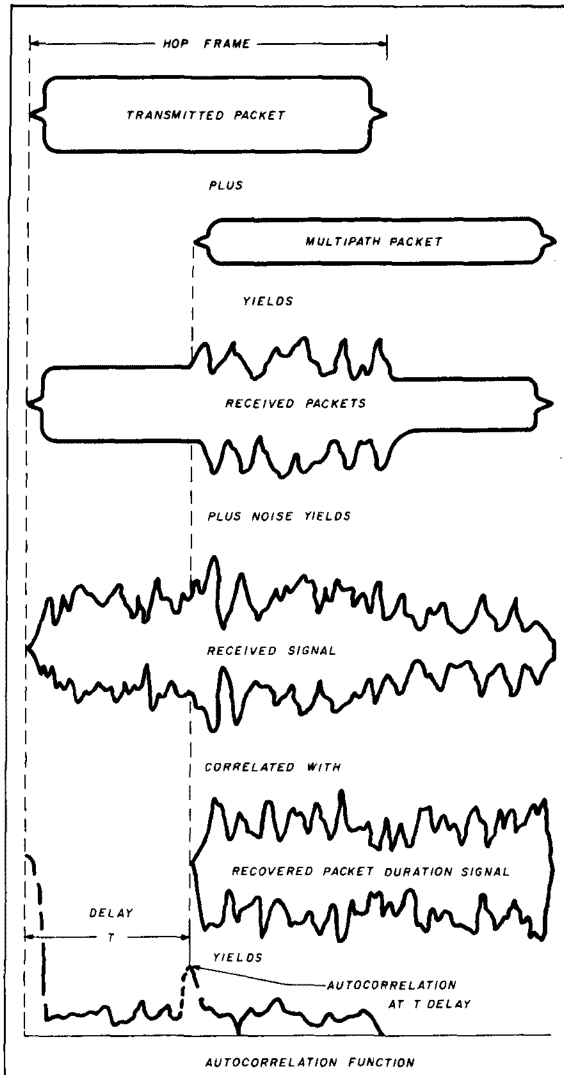


fig. 9. Time relationships in simulated signals.

The principle behind the RCA technique is illustrated in **fig. 10**. Spectral analysis is performed at the output of the power law device (i.e., a frequency doubler or quadrupler). The RMS value of all the distortion products (that is, all spectral lines except the desired carriers) is *directly dependent* on the multipath amplitude and phase and are independent of the data contents. Therefore, no training signal is required to determine the channel quality.

When a biphase PSK signal, for example, is received, the frequency doppler will change a 180 degree

phase shift to 360 degrees. The data now appears as 0 degrees or 360 degrees, which are equivalent angles. The data modulation has effectively been stripped off and converted to a carrier component; this is precisely how many coherent receivers track phase modulated data. However, the carrier component may not be the only spectral energy that remains after doubling. Any distortion products, including the intersymbol interference produced by the multipath delay, produce modulation sidebands. *A straightforward measurement of carrier to sideband energy is then used to determine the amount of multipath.*

When QPSK signals are transmitted, the standard carrier recovery technique in coherent receivers quadruples the signal, thereby eliminating the data modulation leaving only an unmodulated carrier signal. Tracking is then accomplished on what has essentially become a reconstituted carrier.

NOTES:
1. DATA MODULATION IS REMOVED FROM THE BPSK SIGNAL WHEN DOUBLED, THE QPSK SIGNAL WHEN QUADRUPLED, AND FROM THE MSK WHEN DOUBLED.
2. CARRIER, DATA CLOCK AND DISTORTION PRODUCTS REMAIN.
3. DISTORTION PRODUCTS (RMS VALUE) ARE DIRECTLY DEPENDENT ON MULTIPATH AMPLITUDE AND PHASE.
4. BASED ON RCA PATENT 4317206, (ON LINE QUALITY MONITORING).

fig. 10. Distortion measurement approach.

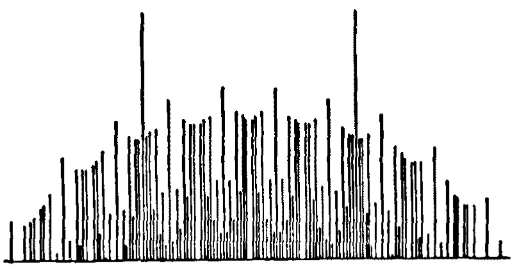


fig. 11. Spectral density MSK after squaring.

When MSK signals are sent, the standard approach to carrier extraction is to double the frequency in the squaring circuit to produce two carriers separated by the clock frequency. This is illustrated for continuous phase FSK in **fig. 11.** In the case of MSK, after squaring, one half the power is distributed between the two carriers. It is standard practice to phaselock a loop to one of the carriers or use the data clock to collapse both carriers into a single component and then phaselock for coherent demodulation of the data. Channel distortion results in less energy in the carrier components and more in the sidebands. Again, the ratio of carrier energy to sideband energy provides an estimate of the channel quality measurement.

The tap weight calculator uses the Channel Quality (CQ) measurements to progressively reduce the uncertainty of the phase and amplitude weights applied to each delay line tap. The CQ is the ratio of the carrier power to the RMS value of all other frequency components after frequency doubling. In the case of MSK, the power in both of the reconstructed carriers is added to achieve a higher CQ and therefore a better measurement. As the tap weights are changed, the CQ value will change dramatically and reach a maximum when best equalization has been achieved.

The simulation example shown in **fig. 12** illustrates rapid tap weight convergence for an assumed multipath amplitude of 0.9 and phase angle of 169 degrees. While seven complete iterations are shown to illustrate the sequence, after the first iteration the phase error is only 11 degrees and the amplitude error 0.4 volts. For purposes of achieving minimum bit error rate performance, this represents complete equalization. Since there is no significant change in the results at the end of the second iteration, the convergence procedure would be halted. If the CQ is high enough, depending on noise and the remaining multipath level, the convergence procedure may actually be halted after the first iteration.

fig. 12. Simulation example of equalizer convergence.

data parameters:		multipath parameters:		equalizer parameters:	
sequence	1	delay (bits)	3.00	tap location	3.000
frame length	64 bits	amplitude	0.90	threshold	1.000
samples/bit	8	phase	169.00	input SNR (dB)	5

iteration	amplitude	phase	amplitude error	phase error	channel quality	data errors
0	0.000	0.0	−0.900	−169.0	25.6	9
1	0.500	180.0	−0.400	11.0	16.2	0
2	0.500	180.0	−0.400	11.0	16.2	0
3	0.500	157.5	−0.400	−11.5	16.4	0
4	0.830	157.5	−0.070	−11.5	212.3	0
5	0.995	157.5	0.095	−11.5	191.5	0

Acquisition process. Because the equalizer does not require training signals, a frequency hopping receiver can achieve initial acquisition and convergence of its equalizer without requiring a cooperative transmission. **Figure 13** shows a transmitter hopping at a normal rate. The receiver is trying to acquire and determine the tap locations, but this is not apparent to an outside observer. Initial acquisition is accomplished by letting the receiver dwell twice as long on each frequency compared to the transmitter. This means that the receiver will miss alternate frequencies; but, at the same time, it will include in its output data the desired signal as well as the multipath signal. As shown in **fig. 13**, the transmitter hops through six frequencies, but the receiver only sees three of those frequencies. Once acquired, the receiver commences hopping at the same rate as the transmitter. Adding a second receiver can provide coverage of the missed alternate frequencies. The additional expense of a second receiver and a frequency synthesizer are not warranted in most applications.

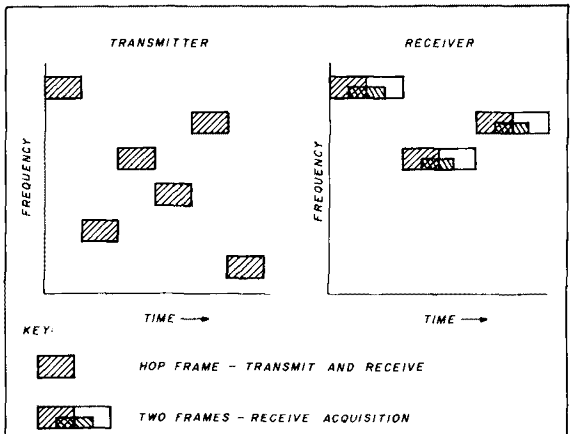


fig. 13. Acquisition frequency hop patterns.

multipath equalizer summary

When frequency hopping is employed to achieve protection from jamming, the problems posed by multipath distortion on HF radio links increase. The novel approach discussed here illustrates what can be done with digital processing. It is well matched to HF hopping systems and supports a very robust form of MSK modulation with efficient error correction coding and is readily implemented with today's technology.

digital SSB tuning

Another example of the successful application of digital techniques to HF radio is the problem of automating SSB tuning. **Figure 14** illustrates a block diagram of a typical digital HF radio. The digital sec-

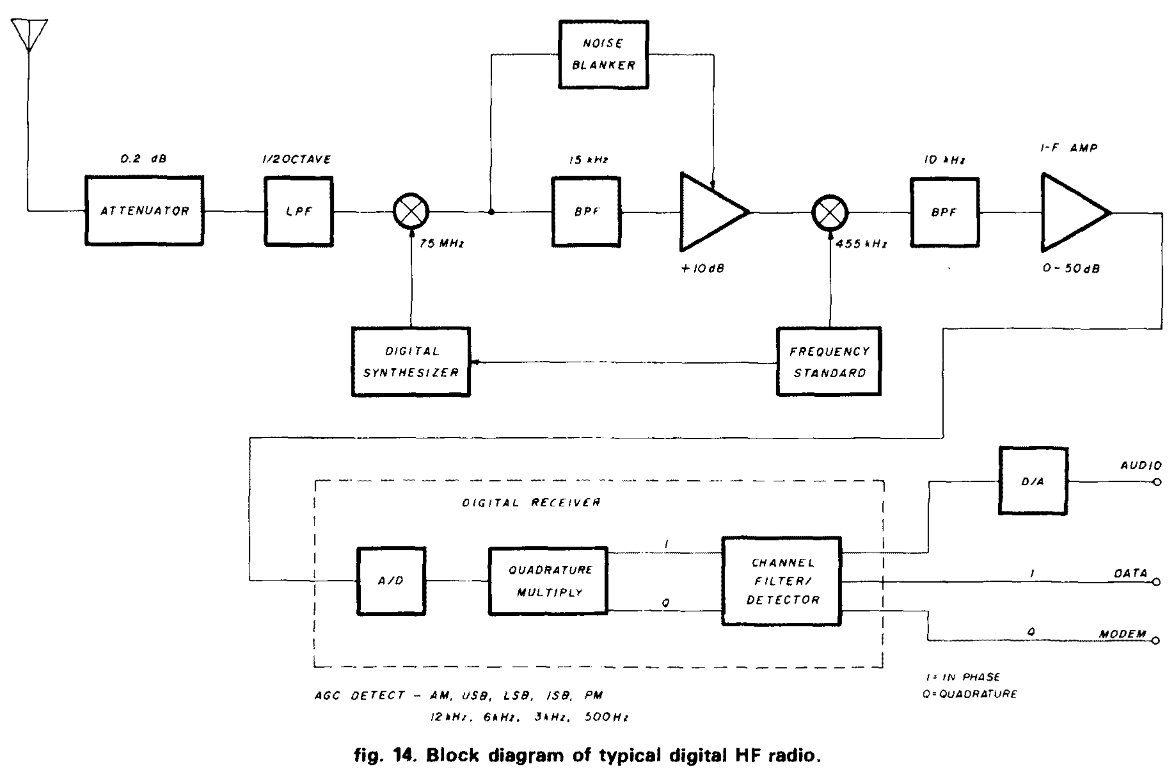


fig. 14. Block diagram of typical digital HF radio.

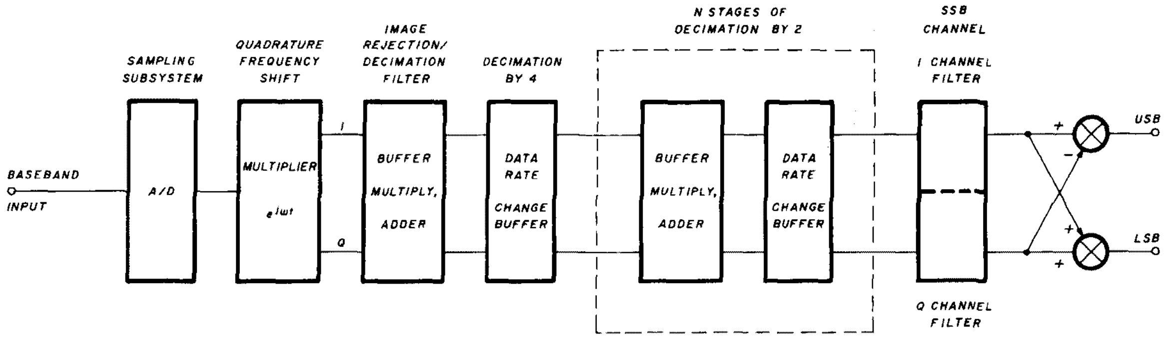


fig. 15. Digital receiver processing (SSB model).

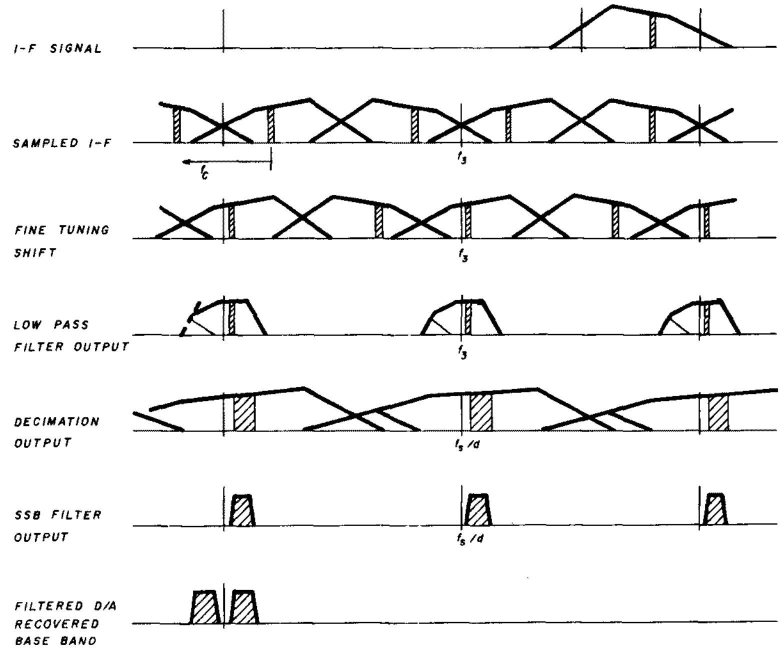


fig. 16. Digital receiver process spectra.

tion of the receiver is indicated within the dotted lines. **Figure 15** shows, again in block diagram format, how SSB detection is performed. **Figure 16** illustrates what signals occur during processing.

The method shown in fig. **17** can be used to develop a software solution to automatic SSB tuning. The voice bandwidth signal is transformed into a voice power spectrum, which is analyzed to detect harmonic relations and complex correlations of the voice spectrum. Digital signal processing then determines the amount of frequency offset relative to a hypothetical center frequency, and a digital signal then tunes the frequency synthesizer to correct any offset. This "center frequency" can then be fed to a processor and stored together with the demodulated data.

Alternatively, this process can be used to regenerate the suppressed carrier. If we look at the lowest signal line in **fig. 18**, we see what appears to be pure noise. If enough samples with sufficient bit resolution are collected, it is possible not only to discover the actual suppressed carrier, but also to find 60 cycle hum sidebands. Spectral detail is useful in uniquely identifying specific pieces of radio equipment. Such techniques are known as "fingerprinting."

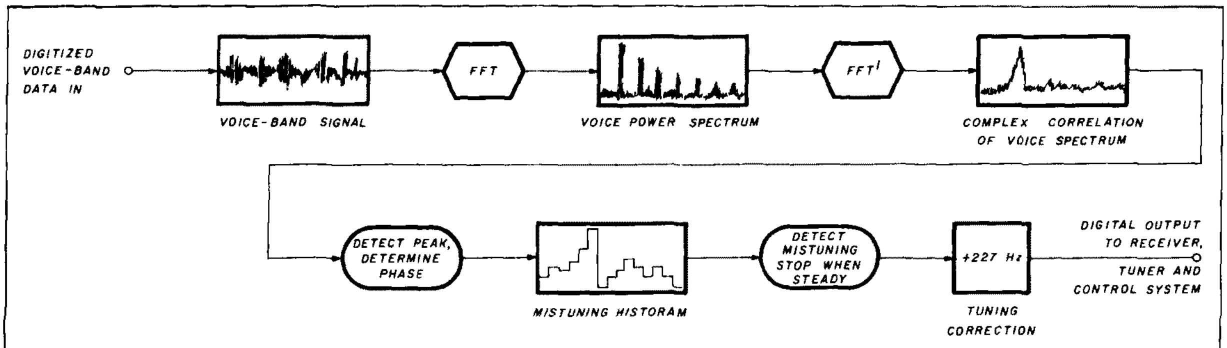


fig. 17. Automatic SSB tuning algorithm overview.

digital waveform generation

Generating analog waveforms from digital signals is another interesting application of this new technology, which allows straight-forward implementation of RCA's "Ampliphase"® system, a modulation scheme used in high-power AM broadcast transmitters.

In its analog form, Ampliphase combines outputs of two individually phased modulated carriers to produce a single AM output signal. Using efficient, nonlinear solid-state amplifiers, it is possible to generate many different forms of modulations — rather than only AM — simply selecting the desired mathematical algorithm. A list of modulations appears in **fig. 19. Figure 20A** shows a digital arrangement that generates AM signals; for explanation, **fig. 20B** shows the amplitude and phase relationships between the two channels. Depending on the phase shift, different modulation and sideband levels are achieved.

The flexibility of Ampliphase is shown in **fig. 21**, which illustrates a method of digitally generating four-channel SSB suppressed carrier modulation. **Figure 22** shows the digital implementation of constant envelope independent SSB suppressed carrier generation. In each case, the hardware stays the same — but a different algorithm is applied.

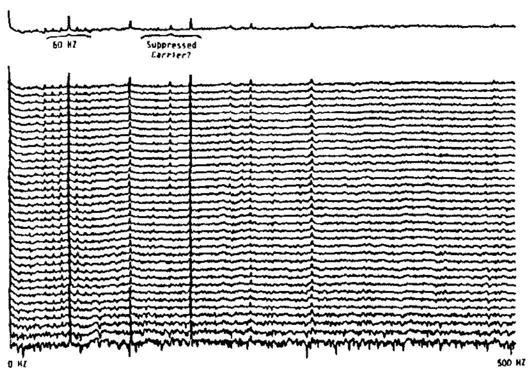


fig. 18. Exponentially-averaged SSB spectra ($\alpha = 0.9$) **mistuned up 100-200 Hz.**

digital filter implementation

Perhaps the most familiar application of digital

fig. 19. RCA's Ampliphase® system generates many different forms of modulation.

modulation:	$x(nT) = Re\,[Z(nT)\,EXP\,(j\omega nT)]$
DSB-AM	$Z(nT) = \frac{1}{2} + \frac{m}{2}\,y(nT)$
DSB-SC	$Z(nT) = y(nT)$
SSB-SC	$Z(nT) = y(nT) \pm j\,\hat{y}\,(nT)$
PM	$Z(nT) = EXP\,[jK\,y(nT)]$
FM	$Z(nT) = EXP\,[jK\,\Sigma\,y(nT)]$
demodulation:	
DSB-AM	*envelope detector* $y(nT) = L + 0.3\,S \approx \sqrt{I^2(nT) + Q^2(nT)}$ *synchronous detector* $y(nT) = Re\,[(I(nT) + jQ(nT))\,EXP\,(j\omega_o nT - \theta_o(nT))]$
DSB-SC	*same synchronous detector as DSB-AM*
SSB-SC	*product detector* $y(nT) = Re\,[I(nT) + jQ(nT))\,EXP\,(j\omega_c nT)]$
PM	*angle detector* $y(nT) = \theta(nT) = ARCTAN\,(Q(nT)/I(nT))$
FM	*discriminator* $y(nT) = \theta(nT) - \theta((n-1)T) \approx \tilde{\theta}$

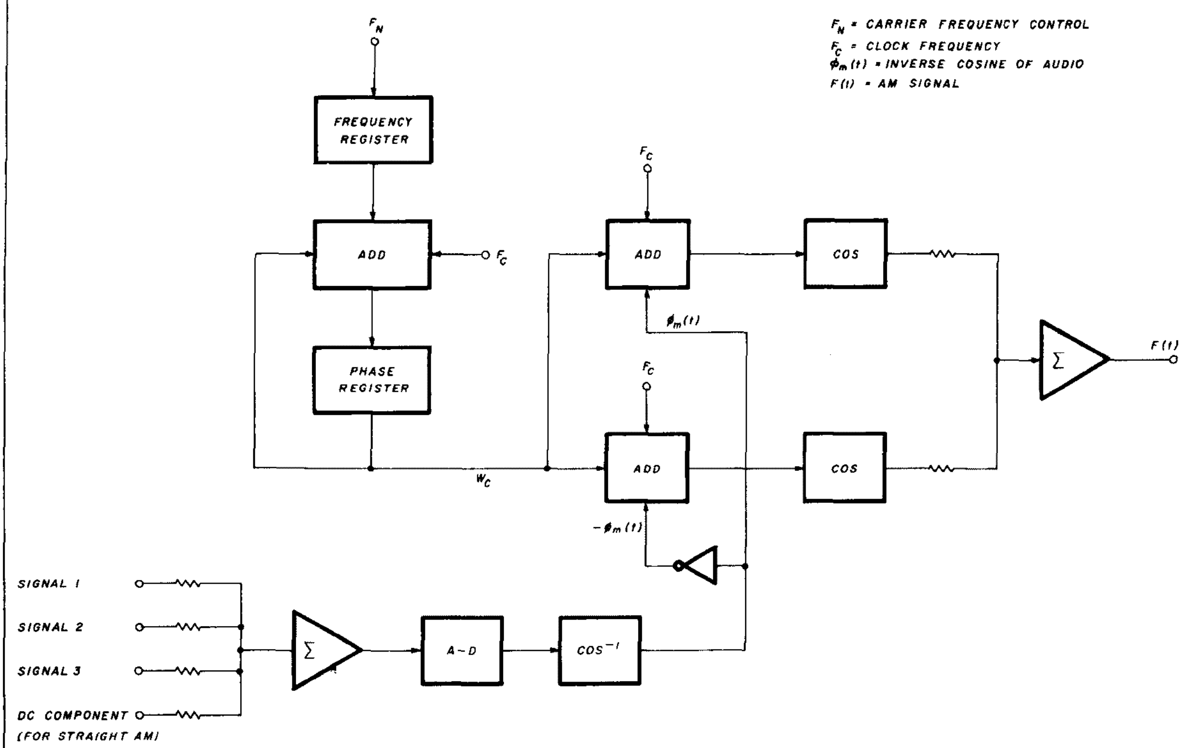


fig. 20A. Ampliphase® AM simulator block diagram.

techniques today is the use of digital filters to replace analog components in audio sections of radios. Often, the selectivity of a digital receiver is determined by the quality of the implementation of its digital filters. These digital filters, typically non-recursive finite impulse response filters, are characterized by no more than sufficient performance, although this deficiency is offset by the simple hardware requirements for implementation. The impulse response of these filters can be computed by using the Parks-McClellan FIR design program. **Figure 23** shows predicted performance data for such filters. The filter algorithms are usually implemented entirely in microcomputer software when time allows, or, alternately, using high speed arithmetic processors to enhance throughput. Special arithmetic logic units can also be employed at the output of a filter to generate either the inverse or the magnitude of the output value, depending upon the filter requirements. Finally, the filter output can be either numeric (digital) or analog via a D-to-A converter.

Figure 24 illustrates the predicted amplitude response for a typical FIR filter.

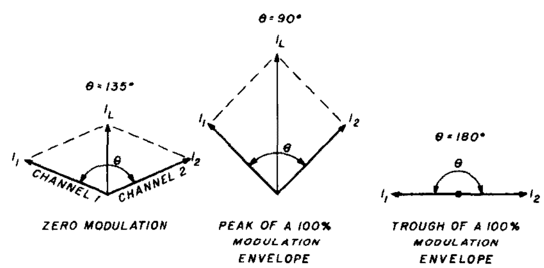


fig. 20B. Vector representation of phases between channels.

frequency synthesizers

Digital frequency synthesizers, common in today's radios, typically have slow frequency switching speeds and variable noise performance. But modern frequency synthesizers can be built with almost infinite frequency resolution and very fast switching speed. (The typical design trade-off is noise sideband performance vs. switching speed.) Analog PLL frequency synthesizers are typically limited in their switching speed by the loop bandwidth of the system and hence offer only limited resolution at high speed.

The best approach to building fast frequency synthesizers with fine resolution is to combine either a wideband analog loop with a digital direct frequency synthesizer or a wideband analog loop with a fractional

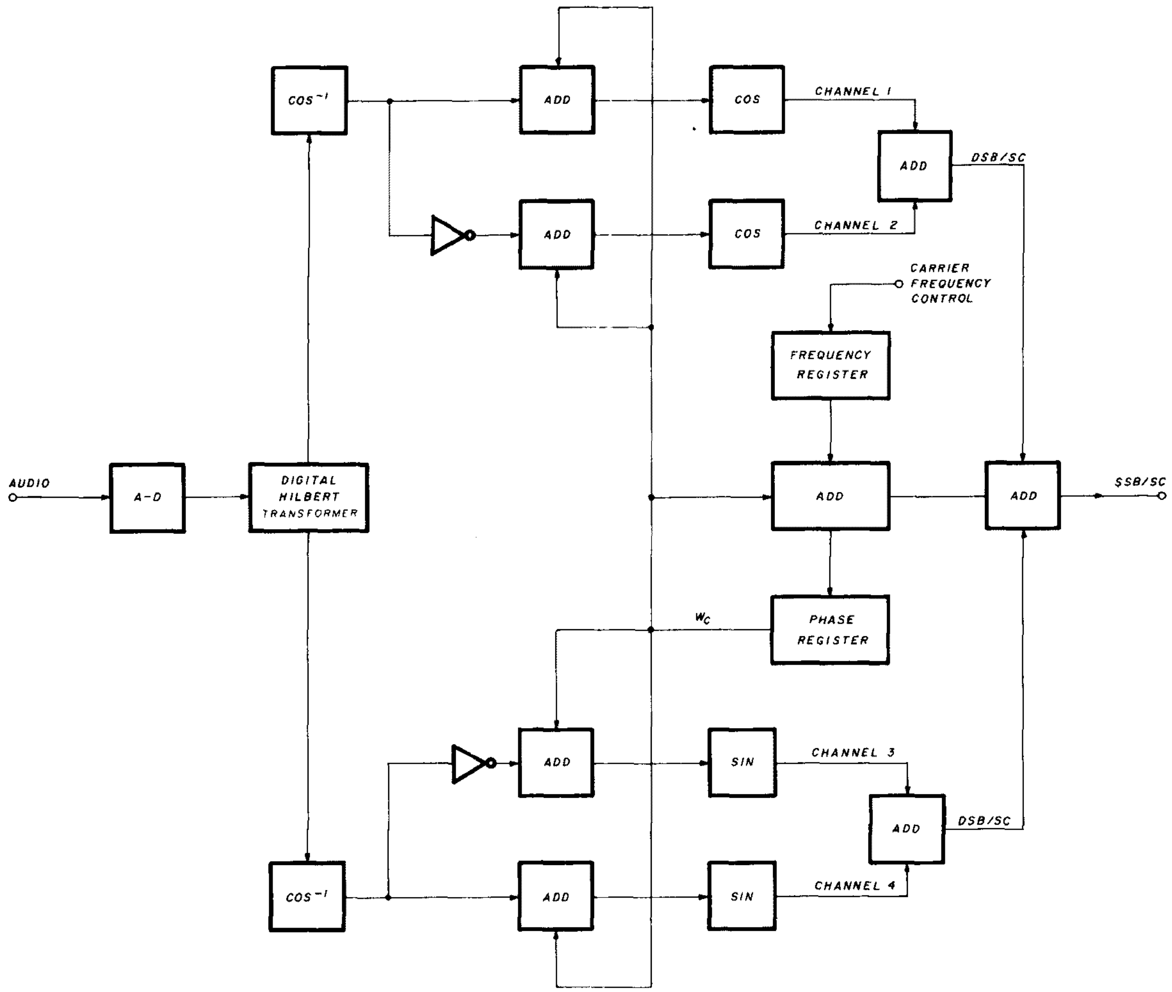


fig. 21. Four-channel SSB/SC simulator block diagram.

divider and synthesizer. **Figure 25** shows one implementation of the former. It is possible to get switching speeds of several hundred microseconds and excellent phase noise at the same time using this technique.

antenna coupler technology

One interesting item not often considered part of the radio is the antenna coupler. Antenna couplers now used provide good RF power transfer between the radios and the antenna. At a single frequency their reliability is not a serious problem. New couplers, with a capability for fast frequency hopping, require new technology to provide fast frequency changes and extended lifetimes – i.e., a predictable period of use without wear-out or breakdown. Because couplers must also be driven by the digital processing system "brain" of the radio, they require a digital interface and, usually, their own processor and training scheme as well.

available hardware

Antenna couplers presently used on aircraft adapt to the driving point impedance peculiarities of their respective antennas. Though tuning methods may differ in detail (depending on the matching network configuration), tuning always involves the use of motor-tuned and relay-switched reactive elements driven toward 50-ohm convergence by an error signal from magnitude and phase discriminators. Tuning accuracies are quite good (VSWR 1.3:1), and coupling efficiencies range between 40 and 85 percent depending on antenna type and frequency of operation. The disadvantage of this class of antenna coupler, however,

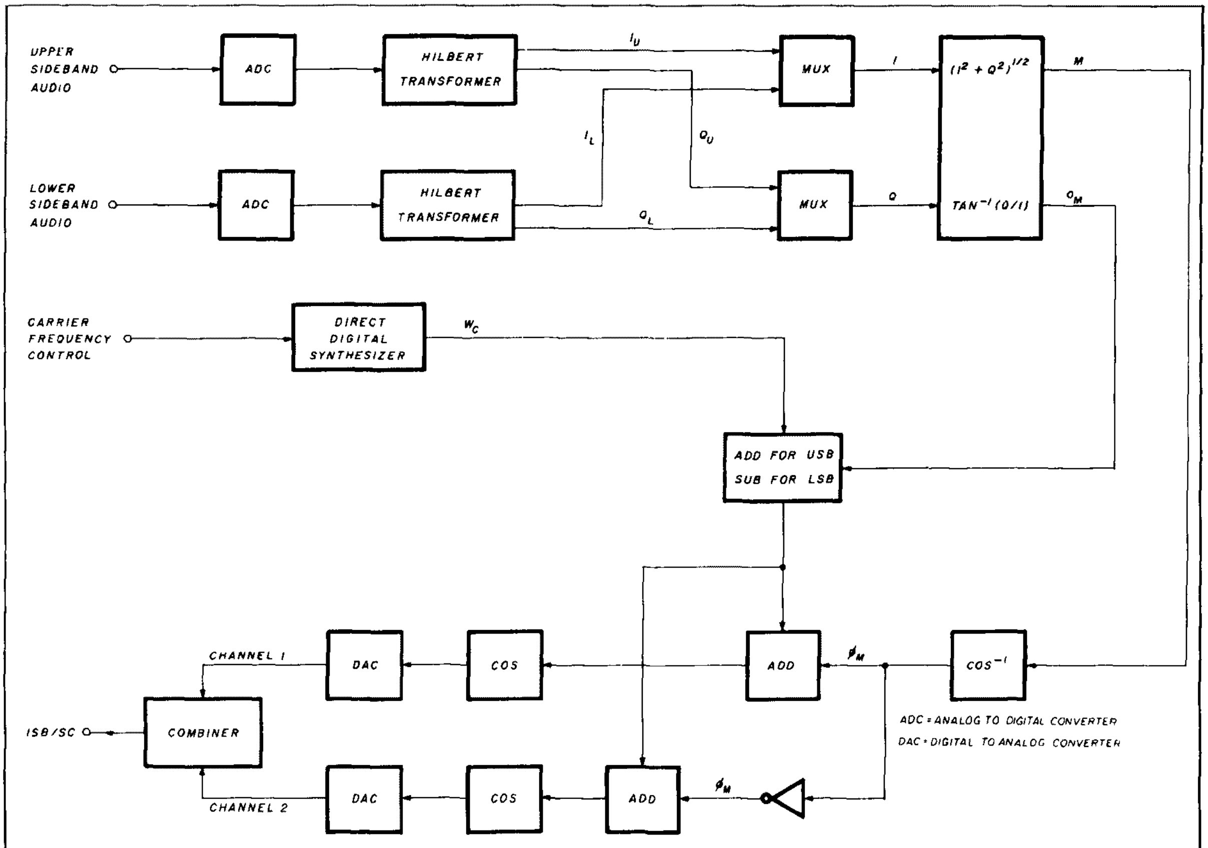


fig. 22. Digital constant envelope ISB/SC simulation block diagram.

fig. 23. Predicted performance data for digital filters in audio applications.

filter function	input sample rate	output sample rate	2-sided passband bandwidth	ultimate rejection	2-sided BW at ultimate rejection	shape factor	impulse response length (μsec)
FM receive	100 KSPS	50 KSPS	25 kHz	>80 dB	36 kHz	1.4:1	62
AM receive	100 KSPS	25 KSPS	12 kHz	>80 dB	17.2 kHz	1.4:1	124
AM receive	100 KSPS	12.5 KSPS	6 kHz	>80 dB	8.8 kHz	1.5:1	248
SSB receive	100 KSPS	12.5 KSPS	2.7 kHz	>80 dB	5.4 kHz	2:1	248

is relatively long tuning time (from 3 to 15 seconds), and a limited lifetime resulting from the failure of electromechanical drives and switches.

Advanced tuning elements. Given the insufficient durability of electromechanical components, the future of tuned couplers with both fast-tuning times and operational reliability will depend on the use of electronic tuning and solid-state devices. Of the solid-state devices that could be considered for this application, two have been given varying degrees of attention: the saturable reactor and the PIN diode.

A saturable reactor is an RF inductor wound on a suitable ferrite core whose permeability is varied by an *orthogonal or parallel* DC excited magnetic field. Inductance variations of 4:1 or greater have been obtained with this kind of device, which can be configured as a memory element by the inclusion of a permanent magnet bias. The main difficulties associated with the use of the saturable reactor are:

- Response time is slow due to excitation time constants.
- Large ferrite volumes are required to overcome the inherent nonlinearity and heat dissipation problems.

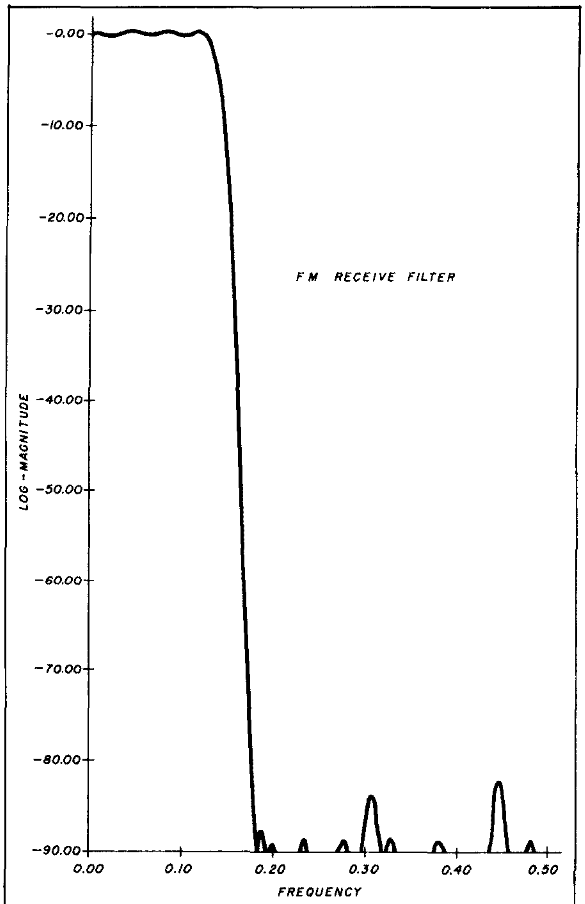


fig. 24. Predicted amplitude response for a typical FIR filter.

- The ferrite material is temperature-sensitive.
- Permanent (irreversible) structural changes can take place at some specific excitation levels.

The device may find application in coupler schemes where a limited degree of adaptive tuning over a narrow frequency range is desirable.

The PIN diode, a solid-state device, offers many interesting possiblities for switching if optimized for operation in the HF frequency range. The keystone of advanced HF Coupler design at RCA, its use permits reactive elements of the matching networks to be varied by switching discrete values of inductors and capacitors in a digital manner. Such a configuration is readily adaptable to a microprocessor control interface. The diode parameters most critical for this application are forward bias resistance, reverse-bias breakdown voltage, minority carrier lifetime, and power-handling capability.

The reverse breakdown voltage of the PIN diodes in the HF frequency range must be high enough to withstand the peak RF voltages developed at the high Q end of capacitive antennas. As an example, an aircraft probe antenna with a Q of 500 at 2 MHz, when matched may develop an RF voltage at the base of 15 kV or more with 1 kW input, depending on the losses encountered in the matching network. Thus a PIN diode, or rather a PIN diode package, must be able to withstand at least half that voltage.

Agile antenna couplers using PIN diodes which quickly respond by microprocessor control can meet today's needs of high hopping rates. Frequency

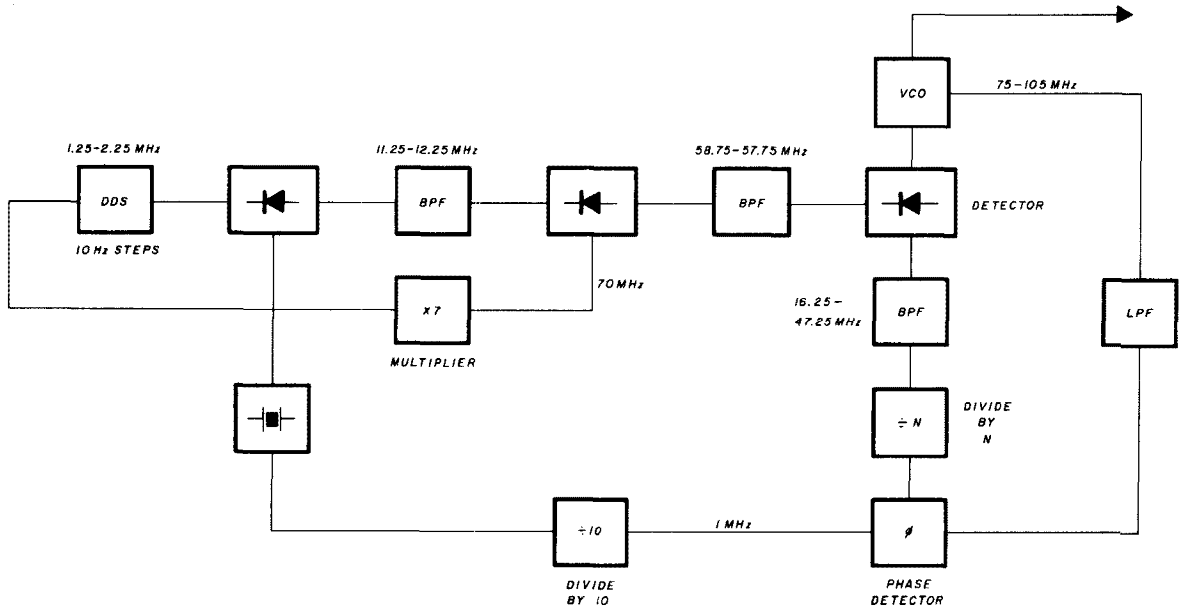


fig. 25. High-speed synthesizer.

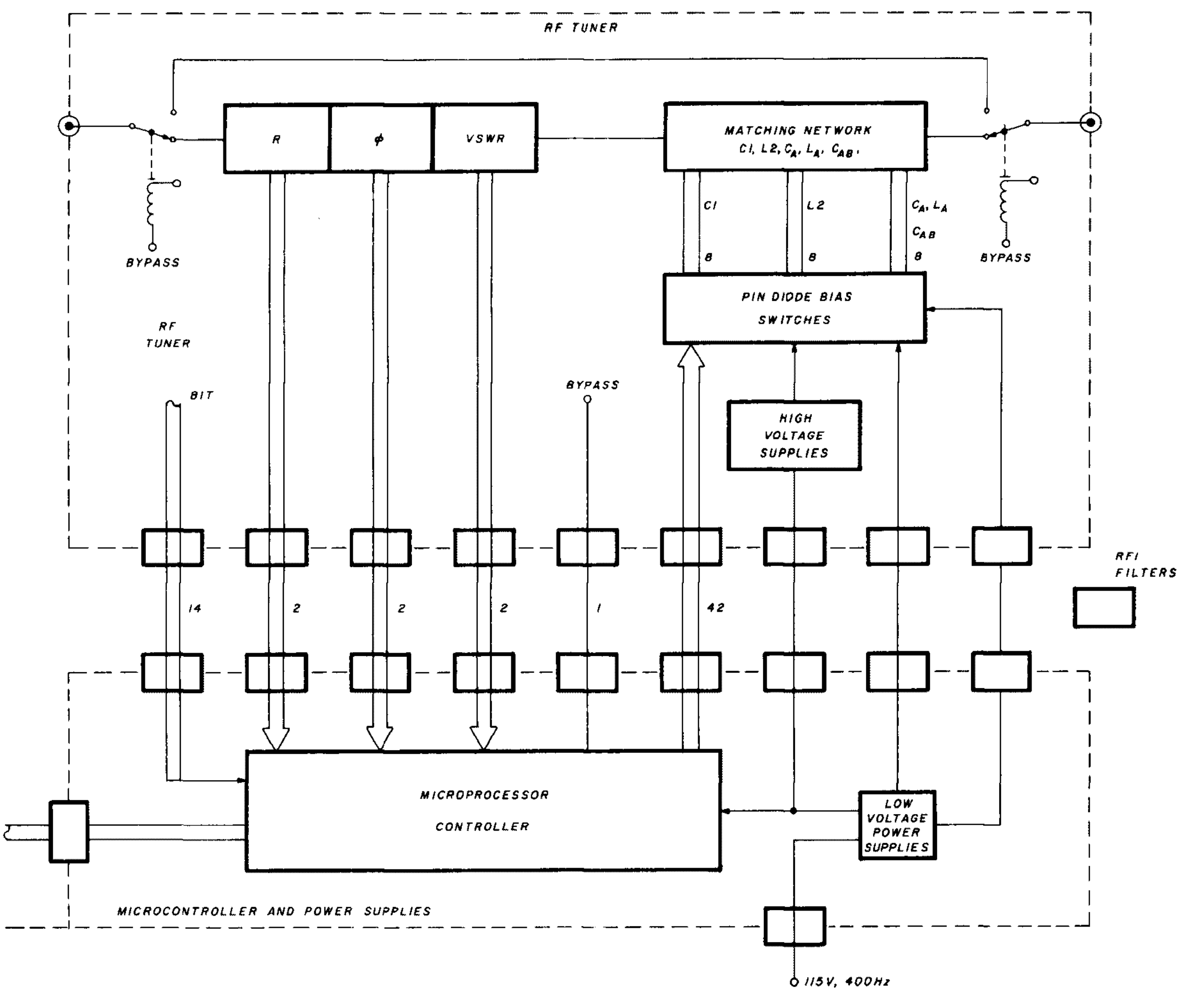


fig. 26. HF agile antenna coupler system block diagram.

changes can be made in microseconds, thus obsoleting older techniques. Advantages of agile antenna couplers with digital processing are high speed, reliability, and adaptability. Through self-test techniques and learning and storing of new antenna characteristics, the coupler can adapt to changes and continue to implement the best impedance match, resulting in good power transfer to the radiating antenna. **Figure 26** shows the system block diagram of the HF agile antenna coupler. A switching speed in the order of 200 microseconds is possible.

summary

This article describes requirements and techniques used in the design of modern digital HF radio which fulfill the needs of advanced RF communication systems. Most of the advanced techniques were developed to satisfy digital data transmission and frequency hopping requirements. Digital implementation of many of the system functions, required for both hopping and non-hopping modes, avoids some of the problems analog linear devices introduce. Digital techniques may ultimately prove to be more cost effective as well. The use of microprocessing and signal processing devices provides greater flexibility of the HF transceiver and allows its use in an integrated system.

acknowledgement

Much of the work described in this article derives from the efforts of the Radio Systems' Engineering section at RCA's Government Communications Systems, Camden, New Jersey, without whose help this article could not have been completed. Among the major contributors to this article were Robert M. Lisowski, John B. McMackin, David A. Miller, Edward

J. Nossen, David P. O'Rourke, and Charles K. Vickers. Additional thanks go to Mr. Miller for very carefully proofing the entire manuscript and supplementing several of the more esoteric terms with examples.

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ham radio

More Details? CHECK—OFF Page 160

a carrier-operated relay
for VHF amplifiers

simple switch operates reliably over wide temperature range

The availability of low-cost hand-held VHF transceivers has increased the demand for external power amplifiers to overcome the inherent power limitations of these radios. In particular, many Amateurs wish to adapt their hand-held units to mobile use and take advantage of the abundance of low voltage DC power available in this application with an add-on amplifier.

Many application notes detailing the construction of such amplifiers are available.[1] A typical design appeared in past editions of the Motorola *RF Data Manual* as EB92A and is built around the MHW-252 hybrid module. The possibility of obtaining 20 dB or more of gain in a single module prompted us to build these amplifiers with the hope of developing 25 watts from the 200 mW provided by an ICOM 2A operating on low power.

The construction of these amplifiers proved to be a near-total disaster for several reasons, and we are by no means encouraging others to follow in our footsteps as far as the amplifier design is concerned. In fact, the latest edition of Motorola's *RF Data Manual* does not list the MHW-252 at all. However, one significant improvement was made in the Motorola design which will undoubtedly prove to be very useful in future power amplifier projects.

The major improvement concerns the COR circuit. In the original Motorola design, shown in **fig. 1**, it can be seen that a 5-pF capacitor is used to couple RF energy to a transistor that drives a mechanical relay. At 146 MHz, 5 pF represents 220 ohms of capacitive reactance. Because the transistor switch is forward biased somewhat during reception, the 5 pF capacitive reactance is primarily the input impedance to the circuit, and thus seriously affects the receiver signal path. In addition, the transistor circuit was found to be extremely unreliable because of base-emitter threshold changes with temperature, especially bothersome in mobile operation during cold mornings. In fact, the slight variation with frequency of the power output of an IC-2A became very noticeable on some mornings, yielding operation on only part of the band.

some problems occurred

Some history of our construction experiences offers comic relief value and should therefore be expounded for completeness. We built three of these amplifiers using the suggested single-sided PC board. As described in the application note, the circuit purportedly used the lead inductance of the relay and printed circuit traces as elements of a harmonic output filter. Measured with a Bird wattmeter, each of our separately constructed units produced a whopping 17-18 watts of output power. This was about 10 watts lower than expected. After a call to the manufacturer we were certain we had erred somehow. The manufacturers' representatives verified the published performance specifications, which we were obviously not meeting. In order to rectify matters, we delved into the theory of the LPF and tried different capacitor values and types. In spite of these changes, output power remained below 20 watts even with different ICOMs, each of which provided the required 200 mW into a 50-ohm load. Suspecting some type of mismatch condition either at the input or output, we proceeded to construct new boards using 50-ohm microstrip for RF connections. Instantly, 30 watts appeared at the Bird wattmeter load upon test. We concluded that the relay lead inductance was itself not enough to mismatch the input and output circuits but that the

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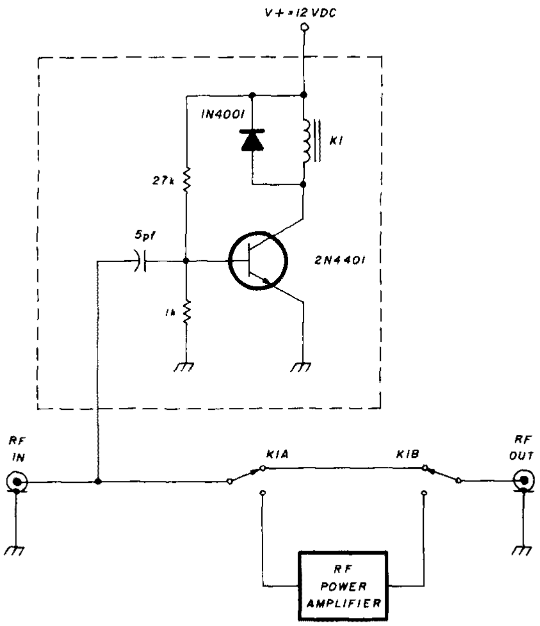

fig. 1. System diagram for 2-meter power amplifier described in Motorola Application Note EB92A. Carrier operated relay shown within dotted line.

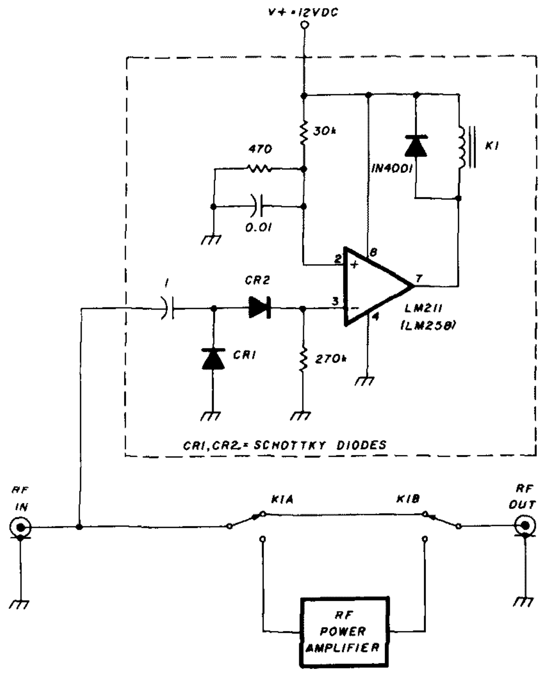

fig. 2. Modified carrier operated relay to provide less receiver loading and operation at power inputs of 100 mW to 10 watts.

circuit traces were. The LPF was also deleted. Other harmonic filters were investigated, and although power loss was not significant, they were not used for the balance of our tests.

In addition to these problems, discussions with the manufacturer provided critical information regarding the input attenuator network. We were told that deleting the network would result in damage to either the ICOM or amplifier because of the adverse interaction between the amplifier and the ICOM resulting from improper source impedance at the ICOM output (on low power only). A 1 to 1.5 dB minimum amount of attenuation is required for the network to reduce the interaction. Fortunately we found this problem by accident before these discussions, but the ICOMs were not damaged. (Our MHW-252 modules were replaced at no charge.)

The process of changing circuit boards to the microstrip type required some work and a bit of magic which we did not possess. In the absence of magic or luck we noted that the MHW-252 leads promptly detach from the module substrate after one soldering/desoldering operation. We assumed we had defective modules, but units from different sources behaved similarly. A honed soldering iron tip and a steady hand allowed us to reattach the leads to the substrate. (We might add that because our construction and design experience is extensive, we cannot wholly be blamed for the aforementioned problems.)

While waiting to overcome some of these difficulties with the circuit, we set out to improve the COR circuit. We decided that the ideal COR circuit should have the following properties:

- as little loading of the receiver signal path as possible (1 pF at 146 MHz)
- reliable operation down to 100 mW
- simple and inexpensive
- ultra-reliable with respect to temperature

COR description

A quick calculation shows that the peak-to-peak voltage of a 100-mW RF signal on 50-ohm line is about 6 volts. This should, in a properly designed circuit, be more than sufficient to guarantee reliable operation. The trick is to make a suitably high impedance switch so as to allow a 1-pF capacitor to provide the coupling. The capacitive reactance of a 1-pF capacitor is approximately 1100 ohms, which presents little loading to the receiver even if the switch input is highly capacitive.

No problems were experienced with the configuration shown in **fig. 2**. CR1 and CR2 are inexpensive Schottky diodes configured as a voltage doubler to extract the peak-to-peak voltage of the RF signal, aside from the voltage drops of the diodes. The bias built up by the action of the diodes keeps them operating primarily in reverse bias where their capacitance is typically 1 pF. This is entirely acceptable for VHF applications and is significantly better than that of low-cost junction diodes. Furthermore, the insignificant storage time of the hot-carrier diodes makes them much more efficient as an RF level detector.

circuit description

The circuit operates as follows. Application of RF power to the input provides a rectified signal at the diode doubler as described. The input impedance of the comparator (or op-amp) and bleed-off resistor is sufficiently high to allow a significant voltage to be developed at pin 2 compared to the bias applied at pin 3 (200 mV). The comparator changes its output state from high to low at this time. Relay K1 is actuated by the near ground potential assumed at output pin 1, provided the required relay current is within the sink capabilities of the comparator output stage. Upon release of input excitation, the voltage at pin 2 decays to a value sufficiently below the bias voltage at a rate determined by circuit capacitance and the 270-kilohm resistor. Subsequently, K1 is de-energized.

The comparator chosen was an LM211 (preferable) or an LM258 (a dual device). Either can sink the 25-30 mA required to operate the OMRON relay used in the Motorola design. In addition these ICs come packaged in a TO-5 metal can package which permitted the pins to be conveniently inserted into the appropriate holes already present in our circuit boards from the Motorola design. (Incidentally the old transistor circuit was removed and tested for 60 Hz response at 120 VAC.) Although double the component count is used compared to the suggested COR circuit of **fig. 1**, the new circuit easily fits on the circuit board and has been used in other power amplifiers.

The circuit, which met all of our requirements, has been in use for several months. If anything, it is too sensitive, as nearby mobile transmitters will sometimes cause the relay to chatter furiously. We noted this effect only at the Dayton Hamfest and consider the inconvenience of minimal consequence.

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ham radio

a PSK telemetry demodulator for OSCAR 10

Get more out of OSCAR with this original design

OSCAR-10, launched by Ariane rocket on June 16, 1983, is the latest transponding satellite for Radio Amateurs. Its period is just under 12 hours and its orbit very elliptical ($\epsilon = 0.6$); this makes it appear quasi-stationary. Much of the time its range exceeds 21,750 miles (35,000 km), which means nearly a hemisphere is within its view, and for many hours on end. The satellite enables Amateurs with modest equipment to communicate at some time or other with stations almost anywhere on earth.[1]

OSCAR-10 (**fig. 1**) carries two linear transponders, at UHF and in L-band. Mode B accepts 70cm uplink signals and has a 2-meter downlink; mode L is 23cm up and 70cm down.

Associated with each mode are two alternative telemetry transmissions: from a general beacon (GB) or an engineering beacon (EB). Of these, the 145.810 MHz general beacon is used predominantly, and will be a familiar sound to most users of the 2-meter Amateur band. The other frequencies are 145.987 MHz (EB), 436.04 MHz (GB) and 436.02 MHz (EB).

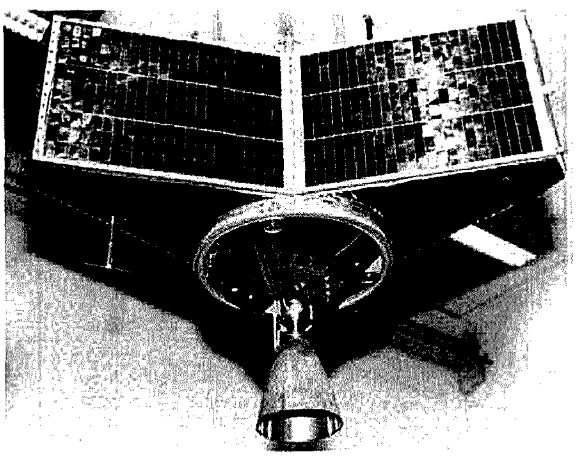

fig. 1. OSCAR 10 (*courtesy AMSAT*).

Transmissions are continuous: on the hour and half-hour UTC there is a 5-minute Morse code bulletin, followed by 20 minutes of phase shift keying (PSK) telemetry. There are two periods of 50 baud RTTY, 5 minutes each on the odd quarter hours.

Telemetry from OSCAR-10 is transmitted in 512-byte blocks, preceded by a four-byte synchronization code (hex 39 15 ED 30) and followed by a two-byte cyclic redundancy check and then a run of about 100-200 padding bytes (hex 50). A byte consists of 8 bits and is transmitted serially, most significant bit first, at a rate of 400 bit/s. So a new block is sent every 12-14 seconds and lasts for 10.3 seconds. The interval before the next block can be used for computer processing of the telemetry.

There are several different kinds of blocks; **Q**, **Y**, and text blocks **K**, **L**, **M**, **N**, are the most common. Their first two characters are always an ASCII identifier, for example M <space>. Line feed and carriage return are in general not used.

K, L, M, and **N Blocks** are plaintext messages, comprising eight lines of 64 ASCII characters. They are at present used for routine communications between command stations.

Y Blocks are entirely ASCII telemetry (**fig. 2**). The first line contains the time (UTC) and AMSAT day number (0 = January 1, 1978). Lines 2 and 3 are command and control status information. Lines 5-8 are 64 selected telemetry values that may be converted using the equations published in the OSCAR-10 operating manual.[2]

As an example, columns 3, 7, 11, and 15 represent temperatures, which decode as T = (N-127)/1.82. The first entries in columns 3 and 7 are the mode B transponder receiver and transmitter temperatures, 16°C and 32°C, respectively.

With the exception of the letter **Q**, **Q Blocks** begin like Y blocks, but lines 5 to 8 contain the full suite of 256 hexadecimal telemetry bytes.[2]

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The schedule of transmissions interleaves ASCII blocks and Q blocks. It repeats approximately every 2¼ minutes.

modulation

The digital information stream at 400 bit/s (called the message) is first differentially encoded such that a 1 is represented by a change in the output data stream (i.e. 01 or 10) while a 0 is denoted by no change (00 or 11). This data is next exclusive-ored with a 400Hz clock, low-pass filtered to restrict its bandwidth (third order Bessel, 560 Hz) and then balance modulated on to the transmitter carrier (**fig. 3**). This modulation is called antipodal phase-shift keying (PSK). Carrier phase is either 0° or 180° according to the data. Because of the low pass filtering there is also some amplitude modulation at bit or clock transitions.

The signal spectrum for random data is shown in **fig. 4**. Note the absence (on average) of a carrier or other line components; these would waste transmitter power.

In the following sections, note the distinction between "message" and "data." The data is the stream which represents the message. Let us use the following notation:

M(n)	n^{th} bit of the original message
D(n)	n^{th} bit of the data, derived from message
S, S(t)	transmitted signal
A(t)	signal amplitude
CLK	the 400 Hz data clock
CAR	carrier
$\oplus$	means EXOR; ($A\oplus A = 0$, $0\oplus 1 = 1$ etc.
$\pm A$	means "either A or its inverse"

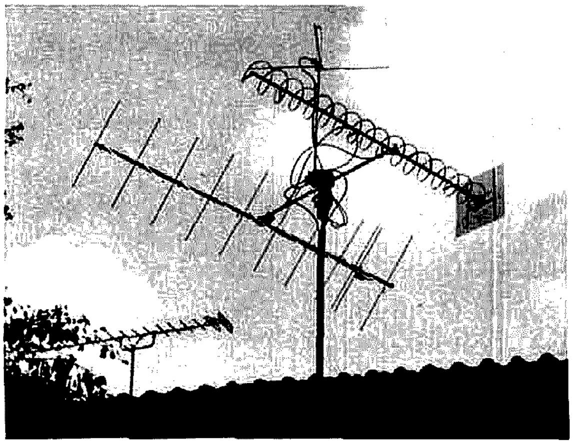
It is useful to remember that if we associate the numeric value +1 to logic 1, and the value −1 to logic 0, then the EXOR operation is equivalent to multiplication, that is: A EXOR B = A × B.

Some features of this modulation scheme that facilitate the demodulation process are the following:

- The signal can be described as $S(t) = A(t)\sin(\omega t)$
- Ignoring amplitude modulation, the signal can also be thought of as the EXOR of data, clock, and carrier: $S = D(n)\oplus CLK\oplus CAR$
- The message stream is related to the data stream by $M(n) = D(n)\oplus D(n-1)$
- Differential message encoding is used to enable a decoder to deal with the unavoidable 180° phase ambiguity in recovered carrier
- All data bits have a mid-bit transition, but not always a transition between bits

demodulation

Essentially this reverses the modulation operations.



Author's helix and cross-Yagi antennas.

The receiver will be set to CW or SSB mode so that the carrier is translated down to audio frequency for input to the decoder (**fig. 5**).

The signal carries negligible information in its amplitude variations, so it may first be limited, which has the great advantage that all subsequent processing can be digital.

First a carrier and clock (denoted by CARR and CLKR) are recovered from the signal (S) and then EXORed with the signal, giving a product (P):

$$P = S\oplus[CARR\oplus CLKR]$$

Provided the local carrier and clock are (excepting possible inversion) replicas of the originals, i.e. CARR = ±CAR and CLKR = ±CLK, this product simplifies to:

$$P = [D(n)\oplus CLK\oplus CAR]\oplus[\pm CARR\oplus \pm CLKR] = \pm D(n)$$

which is the original data. If the signal were noise-free, the data D(n) would be perfectly usable at this point. Noise, however, perforates the bits so mod-bit sampling would lead to random errors. Instead, D(n) is integrated over the bit interval and the resulting accumulation sampled at its end, a process called integrate-and-dump. The system as a whole is a matched filter.[4]

Note that in order to clock the data and time the integration properly, a means must be provided to resolve the CLKR 180° phase ambiguity. (The information to do this is implicit in the signal.)

The message M(n) is next found from ±D(n) by differential decoding. The present data bit is EXORed with the previous data bit. The possible inversion of D(n) is of no consequence, for $-D(n)\oplus -D(n-1)$ and $D(n)\oplus D(n-1)$ are both the same.

The message stream is now available for processing, by either hardware and/or software.

A truly optimum matched filter would take account of the sinusoidal nature of the signal and its amplitude modulation. The hardware required to do this involves deconvolution and is not trivial. The penalty for using a limiter and binary processing is less than 2dB, which in actual on-the-air practice is insignificant.

```
Y  HI, THIS IS AMSAT OSCAR 10                         10:53:03  2308
#034D   #0020   #019E
64  7   0   0   13  225 0

204 0   156 0   199 0   185 126 212 51  152 7   103 47  159 56
0   37  138 31  29  36  140 128 0   0   135 133 111 0   135 13
64  137 141 192 179 170 139 149 118 150 139 13  238 138 148 13
198 140 126 177 198 152 139 0   11  142 141 0   11  137 133 0


Y  HI, THIS IS AMSAT OSCAR 10                         11:08:30  2308
#034D   #0020   #019E
64  7   0   0   13  225 0

204 0   157 0   199 0   185 111 212 51  152 10  103 48  159 56
0   37  139 0   30  36  140 130 10  0   135 146 110 0   135 13
66  137 142 186 177 170 139 96  213 150 140 13  239 138 148 13
198 140 126 10  220 153 139 0   11  142 141 0   11  137 133 0
```

fig. 2. Y blocks are entirely ASCII telemetry. First line includes the day number (2308 indicates April 27, 1984); next two lines contain command and control status information; 64 telemetry values follow.

decoder operation

The receiver should be set to the SSB mode; the normal 2.4 kHz bandwidth is more than adequate, and can with advantage be reduced to around 1 kHz before decoding loss becomes apparent. Because of the decoder's limiter, unnecessarily wide bandwidths reduce performance.

If the signal is tuned in at the middle of the receiver's passband, *then* the carrier frequency will typically be 1500 Hz. The actual frequency is not important, but there is a lower limit caused by the onset of aliasing, when the lower sideband folds around 0 Hz and into itself. This sets in at a carrier frequency of about 500 Hz. The upper limit is set only by the performance of the logic family; if TTL were used, the carrier could well be at 455 kHz — i.e., at an intermediate frequency.

carrier recovery (fig. 6)

A phase locked loop (PLL) cannot be used to extract the carrier directly because there is no carrier line component in the spectrum of an A(t) sin(ωt) signal where A(t) = random data.

However if the signal is subjected to a nonlinear

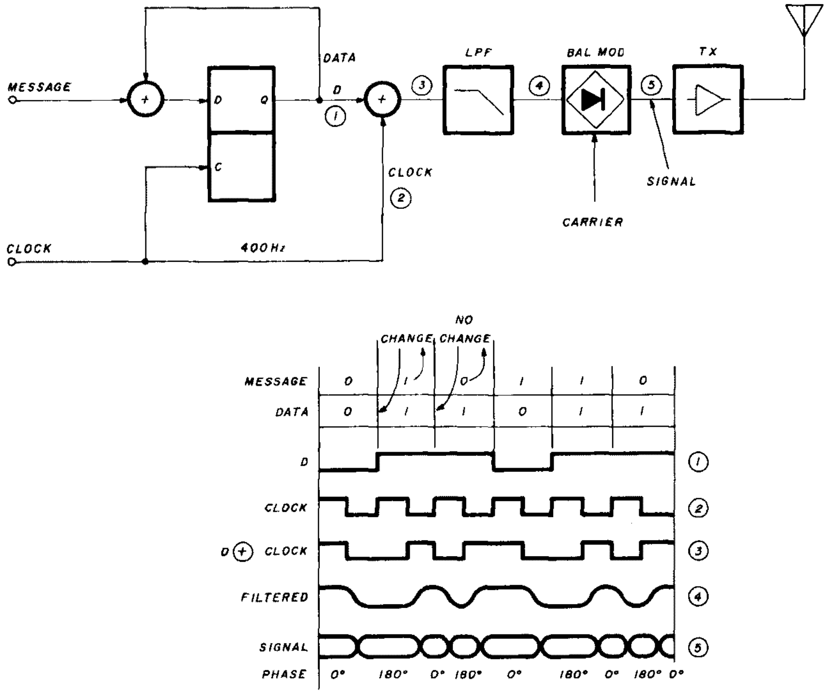


fig. 3. P.S.K. telemetry modulator. Message is first differentially encoded: 0 becomes 00 or 11 (that is, no change in data), while 1 represented by a change (01 or 10). Data is then multiplied (ex-or) with a 400 Hz clock, filtered and modulated on to the carrier.

process such as self-multiplication, the ± is eliminated, and a line component at 2ω is generated. A simple digital way of achieving this is to EXOR the signal with itself delayed by a quarter-cycle. Every zero-crossing of the carrier creates one new cycle of twice the carrier frequency, which can be regenerated with a phase locked loop and followed by a divide-by-two circuit. This division does, however, result in the 180° phase uncertainty previously noted.

The carrier PLL must accomodate receiver frequency instability, noise and changing doppler shift (−250 Hz/hour at 145 MHz, USB. If the loop bandwidth is too small the PLL is difficult to tune in initially, has only a small tracking range, and is generally fussy. If it is wide, with little noise, the loop will hold lock over a wide tuning range, but it will constantly lose lock on noisier signals. So a fixed loop bandwidth will not suit every situation. A few experiments will show what is wanted in practice: the value will lie between 10 and 100 Hz.

clock recovery

This is accomplished in exactly the same way as carrier recovery, by multiplying the data by itself delayed by a quarter-bit. This generates an 800 Hz proto-clock, which is regenerated with PLL and then divided by two. Since the clock frequency is constant, the loop bandwidth can be 1 Hz, even with cassette tape signals. As with the carrier loop, the clock at this stage also has a 180° phase ambiguity.

clock ambiguity resolution (fig. 7)

As long as there are 01's and 10's in the data, which means that the inter-bit transitions will be absent, a second proto-clock of 400 Hz can be generated by EXORing the data with itself delayed by half a bit. Although somewhat sparse (trace 3) this extra clock has the virtues of correct phase, and coherence with the ambiguous clock.

If these two are now EXORed together, the smoothed result is a net high or low. This signal can then be used to invert (or not invert) the ambiguous clock to the correct sense.

In fact the second 400 Hz proto-clock could be used to excite a PLL. However, with the signals encountered in practice, the effective loop gain and bandwidth are caused to vary constantly, which makes the loop rather fragile — though it does work.

A particular feature of the clock and carrier recovery circuits is their aperiodic, digital design, involving no tuned circuits. So their operating frequency can be modified simply by changing the VCO center frequency.

block sync detection

Hex 30, 15, ED, 30 is the pseudo-random sequence generated by the first stage of a five stage shift register having its middle and last outputs EXNORed and fed back to its input, starting off all all 0s.

Using such a feedback shift register, 100% sync detection can be effected by comparing the message stream bit serially with the output of the first stage. If there is a disagreement, the shift register is reset to zero; otherwise it is clock on. If the full sequence is successfully checked the register will reach its last state, which can be detected with a five-bit AND gate and used to set a start-of-block flag. The flag then inhibits the shift register.

byte counting

The block flag releases a byte-block counter. Every eighth count signals that a byte is available, and when the counter reaches 4096 (8 × 512), the block flag is cleared and sync code testing resumes.

outputs

Parallel data output is buffered to TTL level, and consists of an eight-bit byte, positive-going mid-bit strobe, and the block flag. These are brought out to a 20-way PCB connector. Pin-out is compatible with the BBC Acorn microcomputer 6522 user port, which will also provide a 5V supply for the output buffers.

The serializer gives an RS232-type output at 5 volts, 1200 baud, with one start, eight data and many stop bits (50 characters per second). This could be used to drive a printer directly, but the hexadecimal, non ASCII Q blocks will cause unpredictable results — weird characters and reams of waste paper! Using a VDU (video display unit) or the serial port of a computer is tidier.

circuit notes

The complete circuit diagram is shown in **fig. 8**. A printed circuit board is available.[2,5] All other parts may be obtained from Bob Wilson at Radio Kit (Box 411, Greenville, New Hampshire 03048).

A half V_{DD} bias is incorporated to 'float' the op-amps. The 12V supply is not especially critical.

There is no channel filter built into the design because the receiver provides one. The simple limiter U1A* will be effective on a few millivolts of signal. The meter circuit U1B is primarily intended to aid tuning; but with an external switch S1 it can be used to monitor other signals, in particular the state of lock of the two PLL's.

The main PLL, U3, runs at 16 times the carrier frequency. This drives a four-bit shift register, U5, to give

*Note that the author's prepared printed circuit board uses the "IC" designator for integrated circuits. All other figures follow *ham radio* style, designating IC's as U1A, U1B, etc.

— Editor

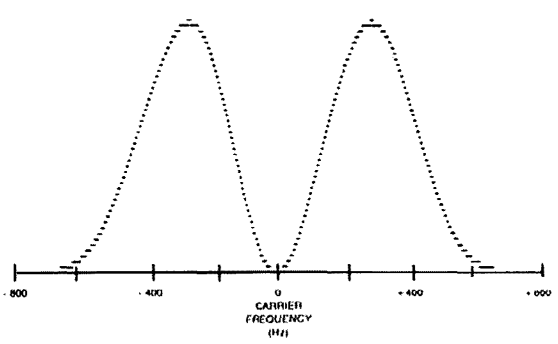


fig. 4. Telemetry power spectral density. Vertical scale is linear. Note the absence of any line components, which would waste power and also make it impossible to lock on to the carrier directly.

Front panel of decoder.

a quarter-cycle delay. The VCO, U3 is followed by a divide by 16 counter, U4, and some logic to generate the local Q:2f signal and the recovered carrier CARR which translates the signal to baseband in U2A. The loop bandwidth is 10 Hz; component values for other bandwidths are shown in the table.

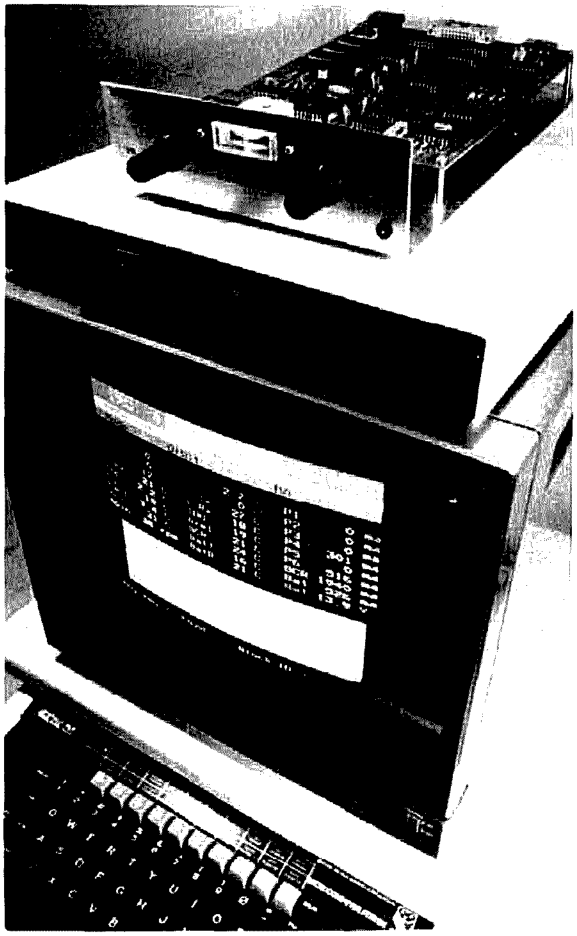
Extraction of the 400 Hz clock is similar to the carrier loop; the VCO, U7, runs at 6400 Hz and has a 1 Hz loop bandwidth. Note the additional quarter-bit delay, U9B, which provides the overall half-bit delay needed for the CLKR ambiguity resolution performed by U10A, U10C, and R13, C14. The clock extraction circuits were originally devised for a UOSAT data demodulator[3] and the integrate-and-dump U1C, U11, and U12A is taken from that source too.

The block sync code detector consists of the feedback shift register code generator U14, U15C and final state (00001) tester, U16A, U16C. The shift register is released when the block bistable, U19A, is clear. Incoming data (bit D(0) is tested against the code generator in EXOR gate U15A. A high state indicates a disagreement, and a CLK pulse resets the shift register. If the shift register reaches its last state, a CLK pulse at AND gate U17D sets the block bistable, UI9A, which also lights an LED. The shift register is reset and inhibited, while the bit/byte counter, U20, is released.

Every eighth count generates a positive, mid-bit byte strobe from the shaper circuit, U18C. This pulse signals external equipment to read the byte in buffer U13, via buffers U21 and U22.

The serializer works as follows. The rest condition has the start bistable, U24B, permanently clocking out "stop" which is cascading through from the shift register DS input, U25, pin 11. When the decoder has a byte ready, byte strobe sets the control bistable U24A pin 1 to "load". This prepares the start bistable U24B to go to 'start' and the shift register U25 to load with bits D(0)-D(7) from the decoder buffer. This happens on the next high-going edge of the 1200 Hz clock, which also clears the load condition. Subsequent 1200 Hz clock pulses shift the data out of the serializer.

Outputs have the following conventions: the parallel data output byte is 1 high, 0 low. D(0) is the least-significant bit. "Block" is high true. Byte strobe is a high going 20 microsecond pulse, and begins mid-bit.



VDU, keyboard, and detector.

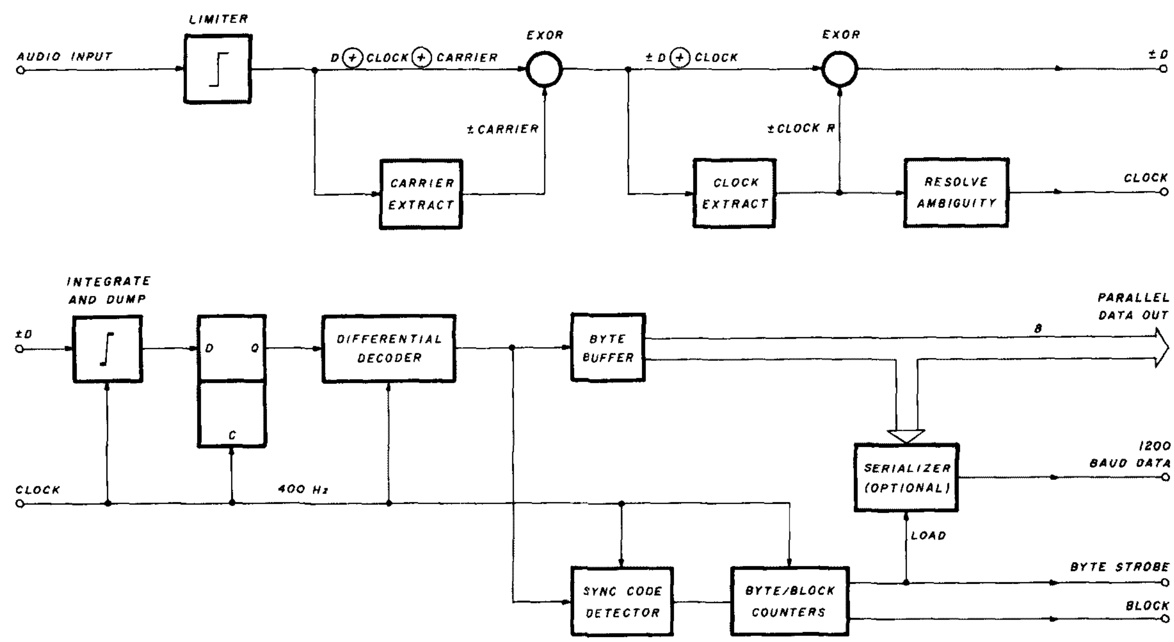


fig. 5. Audio input to the decoder comes from a 2-meter Amateur band receiver. The output is available in parallel or serial form to suit the user's computer.

The serial 1200 baud data format is Start and Data 0 high, Stop and Data 1 low, eight data bits, l.s.b. sent first. The 1200 baud square-wave clock goes low mid-bit. Note that not all data is ASCII in particular the last 256 bytes of Q blocks.

400 bit/s serial data and CLK may be selected as the serial output by changing link D and link C. Square wave CLK goes low mid-bit. The data is *not* in start/stop telegraphy format.

setting up

An audio generator, oscilloscope, and multimeter will be needed. A telemetry data test tape will be invaluable: this can be recorded off-air or obtained via AMSAT-UK[2]. The satellite itself can be used for live testing, but is not always available when you want it. The signal is also noisy, which may confuse matters.

receiver

The first job is to decide what carrier frequency will be used. Trigger the scope at 200 Hz (or 50 Hz if available), tune the receive to OSCAR 10's beacon, and display the audio. Amplitude modulation should be discernible. Experiment with tuning and bandwidth until the signal looks healthy with the mid-bit crossover clear and sharp. Now trigger the scope with the signal estimate and note the carrier frequency, say f_c. Any frequency exceeding 1000 Hz will be satisfactory.

carrier loop

The objective is to set the loop mid-frequency to the measured carrier frequency (f_c) and achieve a total frequency swing (f_{sw}) at the output of the divide-by-16 (TP1) of $f_{sw} = 800$ Hz. This is slightly complicated by the fact that a 4:1 spread in oscillation frequency between different samples of a 4046 VCO is quite typical.

Start with the VCO swing, nominally given by $f_{sw} = \frac{1}{R7 \times C10}$. Apply V_{DD} and then 0V to pin 9 of the PLL chip, U3, measure high and low frequencies at TP, and subtract. If the difference f_d, is within 25 percent of 800 Hz, then all is well. Otherwise, change C10 for the correct swing.

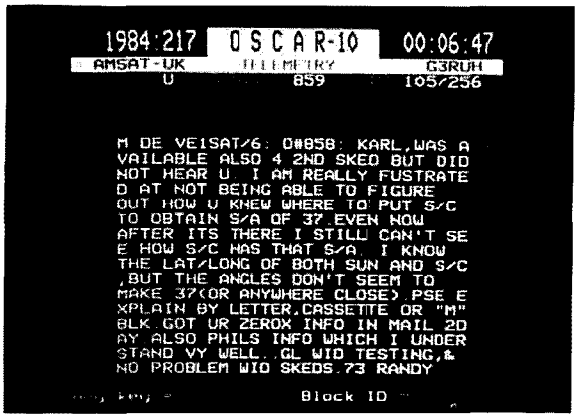
Calculate and note the desired VCO upper frequency, $f_u = f_c + f_d/2$. Again, connect V_{DD} to pin 9 of the 4046. With the main tuning potentiometer, VR_3, at mid-position, adjust trimmer VR_2 to give frequency f_u. If this cannot be achieved, then change C10 and R7 (in inverse proportion to each other, so as to preserve f_d) and start again.

Now inject f_c at the audio input. Check that the loop locks on this signal. The lock meter should indicate to one side. Slightly vary the input frequency and the main tuning control VR_3 and observe the tuning meter center-zero response. The loop should stay in lock over a range of $\pm f_d/2$.

clock loop

In the same way as for the carrier loop, check that the available frequency swing at TP_2 is about 100 Hz in total. If necessary, change C12 to achieve this. Then adjust VR4 so that the mid-frequency is 400 Hz.

Temporarily ground U2A pin 1. Inject 400 Hz at the audio input. The loop should lock up correctly; this will show on the lock meter. Next inject 200 Hz, which simulates data 010101 . . . (message 11111 . . .). Verify that the loop locks again. Examine the ambiguity signal at U10D pin 12. This should be either high or low and should not vary about $V_{DD}/2$. Now examine the CLK signal at U10D pin 10. The low-going edges should coincide with the transitions of the input signal.



Enlarged view of display.

Disturb loop lock a number of times by removing the 200 Hz input signal for a few seconds. Each time this is done the ambiguity signal will assume a random state, but CLK should always resolve itself to the correct phase.

$V_{DD}/2$ supply

Remove the temporary ground from U2A pin 1. Apply receiver random noise to the system input. Connect an analog meter (on VDC) across pins 1 and 2 of the bistable U12A. Adjust the half-supply control VR_1 for zero reading.

testing

Once the system has been adjusted it may be checked out live or with a test tape[2]. The waveforms obtained should be as shown in **figs. 6, 7, and 9**. A number of features of the satellite data make this easier. The padding character hex 50 and <space> both occur in longish bursts. In addition, the sync code tester will obviously not work unless everything else is going properly, and so illumination of the "block" LED once every 14 seconds for 10 seconds provides a quick, comprehensive overall check.

decoding data

The design of the software to decode and display the data is straightforward enough, but it is outside the scope of this article to present it in full.

The computer should examine the block flag until

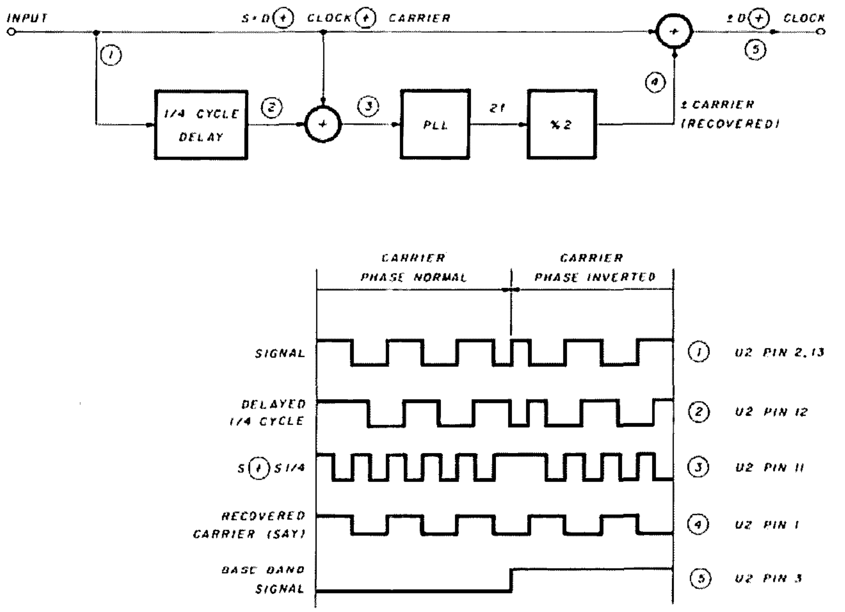


fig. 6. Carrier recovery and conversion to baseband. Clock recovery is identical: simply add 4 to each of the U numbers.

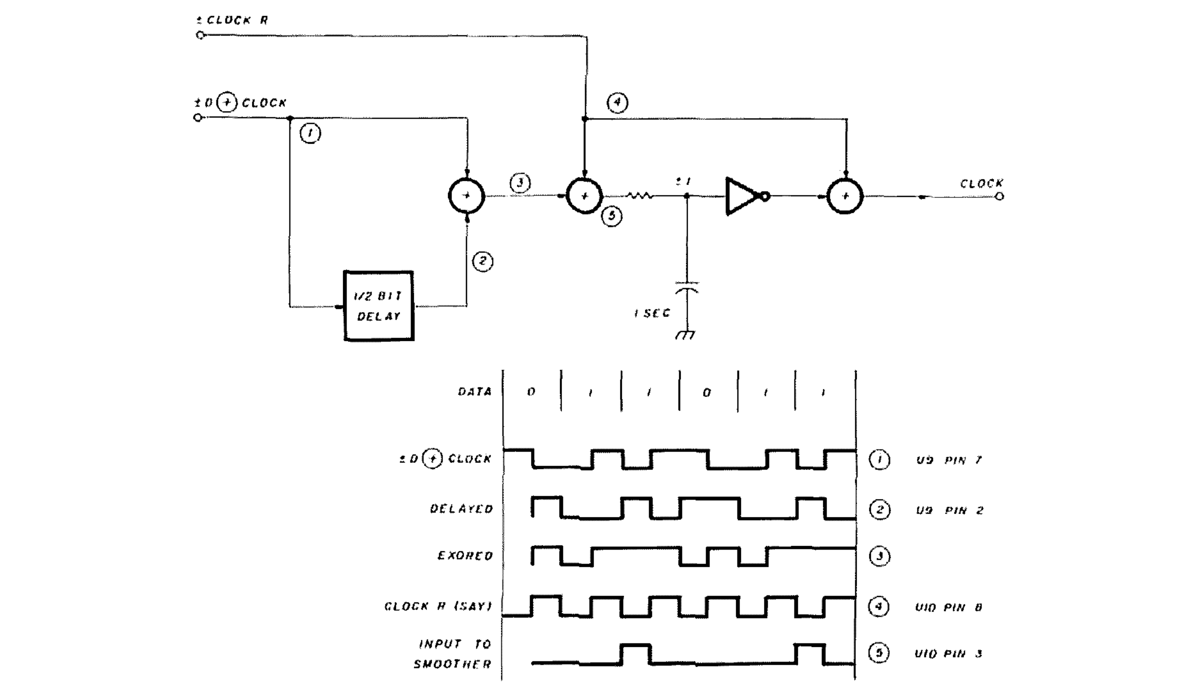
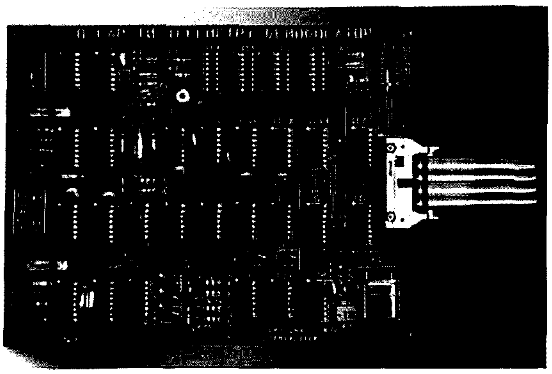


fig. 7. Resolving clock ambiguity: input to the smoother results in a net low on the capacitor, so ex-or inverts *ClkR* to the correct phase, *Clk*. If *Clk* had been correctly phased in the first place, then the capacitor would hold high and *ex-or* would not invert.

it is asserted, then wait for a byte strobe. It should then read in the byte; place it into a 512-byte buffer and await the next strobe. Alternatively, bits may read in serially and packed away.

When all 512 bytes have been read, decoding can begin. In real-time there are 4 seconds in which to do this. Check that the first two bytes are recognizable identifiers, e.g. Q <space>. Then all that remains is to pick out the items of interest such as volts, amperes and temperatures and to display them on a printer or screen in an appropriate format.

Alternatively it is possible to dump the lot, or selected bytes, to storage for later processing, perhaps to monitor specific parameters or to plot graphs.



Decoder board with ribbon connector and decoder.

performance

A useful indicator of performance is given by the bit error rate. If we define a reasonable rate as less than 1 in 10,000 bits, i.e. an average of one error every other block, the theoretical channel signal-to-noise ratio S/N should be 2.4dB in 1600 Hz bandwidth. Allowing for the signal amplitude modulation and the limiter, the practical figure is actually about 6.2dB, peak signal power to noise power, or 2:1 in voltage. With care this can be verified experimentally — the signal sounds and looks pretty ragged.

An S/N of 6.2dB is represented in the lab by the surprisingly small figure of 52 nanovolts (nV) (− 133dBm) at the input of a receiver having a 3dB noise figure. Now, the 2m general beacon transmits about 1W (+ 30dBm); the space loss over a 24,850 mile (40,000km) path is 168dB, so the received signal at a unit gain antenna is roughly − 138dBm. Thus an antenna gain of 138 − 133 = + 5dBi is needed, plus a margin for fading, cable losses, wider bandwidth, higher receiver noise figure and so on.

In practice this means that for satisfactory recep-

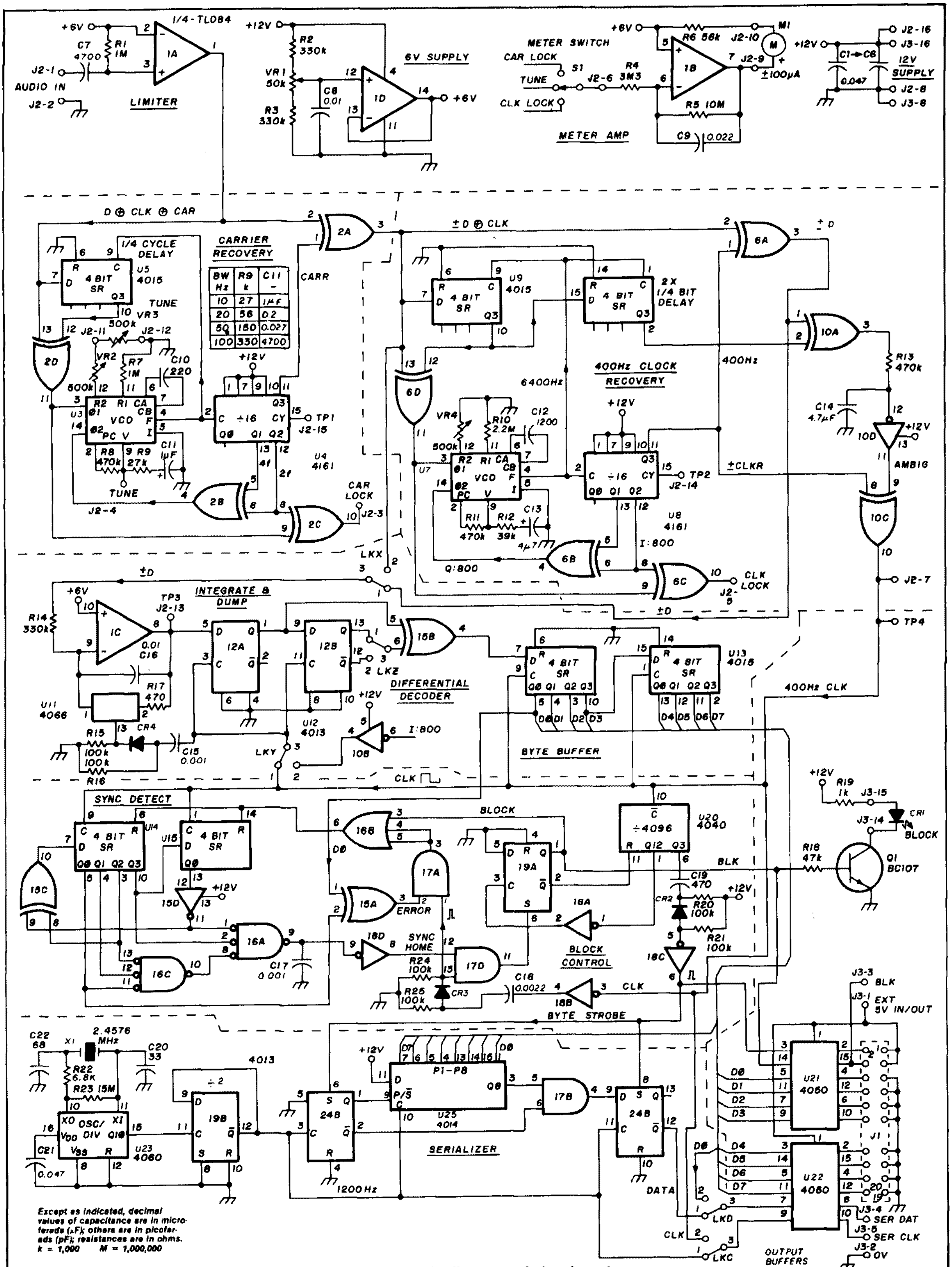


fig. 8. Circuit diagram of the decoder.

Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

OSCAR-10 PSK demodulator parts list.

capacitors (all 16 volt rates)

C1-C6,C21	0.047 μF
C7	4700 pF
C8	0.01 μF
C9	0.022 μF
C10 — see text	220 pF 5 percent,*
C11 — see text	1 μF tantalum
C12 — see text	1200 pF 5 percent*
C13,C14	4.7 μF tantalum
C15,C17	0.001 μF
C16	0.01 μF**
C18	0.0022 μF
C19	470 pF
C20	33 pF
C22	68 pF

*polystyrene
**good polyester

CMOS integrated circuits

U1	TL084	quad op amp
U2,6,10,15	4070	quad EXOR
U3,7	4046	PLL
U4,8	4161	divide-by-16
U5,9,13,14	4015	quad-4 bit SR
U11	4066	quad switch
U12,19,24	4013	dual-D type
U16	4075	triple-3 OR
U17	4081	quad-2 AND
U18	4069	hex inverter
U20	4040	12-bit divider
U21,22	4050	hex buffer
U23	4060	Osc/14-bit divider
U25	4014	8-bit SR

potentiometers

VR1	50k preset, cermet. Spectrol 62
VR2,4	500k preset, cermet. Spectrol 43
VR3	500k linear, carbon

resistors (all 5 percent)

R1, R7	1M
R2,R3,R14	330k
R4	3.3M
R5	10M
R6	56k
R8,R11,R13	470k
R9 — see text	27k
R10	2.2M
R12	39k
R15,R16,R20 R21,R24,R25	100k
R17	470
R18	47k
R19	1k
R22	6.8k
R23	15M

semiconductors

CR1	LED 10 mA red
CR2,3,4	1N4148 (or equivalent)
Q1	BC107 (or equivalent)

miscellaneous

M1	± 100 μA center zero meter
X1	2.4576 MHz crystal HC33/U size
J1	20-way PCB header for IDC connector
J2,3	16-pin DIL socket
S1	1P3T (1 pole, 3 position) switch
TP1-4	test points

Note: The meter, VR3, CR1, and the switch, S1, are not mounted on the board. Links LKC, D, X, Y, and Z are made from hook-up wire. PC board designators have been left in British style.

tion a modest Yagi or equivalent is needed, pointed at the satellite.

It is worth noting that it is typical of optimal demodulators that they exhibit a marked performance threshold effect. In our "6.2dB" example above, a reduction in the S/N of only 1dB results in a dramatic tenfold error rate increase. This is most apparent where there is a rapid fading (usually induced by the satellite's 40 rev/min spin): what appears to be a healthy signal actually results in bursts of errors at S/N minima. Spin fading occurs most strongly a few hours each side of apogee, when the spacecraft's antennas are not pointing directly towards Earth.

Because of the differential decoding scheme, a single bit error leaving the integrate-and-dump section results in two adjacent bit errors at the system output. This should be remembered if any software error checking is to be attempted.

a further decoding method

Finally, there is another method of decoding the signals. There is a distinctive relationship between the message bits (as opposed to data bits) and the encoded stream. Each message[1] results in a D⊕CLK signal with missing inter-bit transitions, whereas a message 0 does not (**see fig. 3**).

So an alternative decoding method is to treat D⊕CLK as a stream of 800 bit/s half-bits, grouped in pairs. Two similar successive half-bits are decoded to a logic 1 output, and two differing half-bits to a 0.

This can be implemented most simply by feeding the integrator with D⊕CLK, clocking the intergrate-and-dump and differential decoder with $\overline{\text{I:800}}$, and inverting the data output sense! Links X, Y and Z are provided to enable experimenters to evaluate this.

The error properties of this arrangement are interesting. Because the signal energy per dump decision has halved, the half-bits' intrinsic error rate is much higher than a whole bit's, but it is now possible for single message bits only to be corrupted.

The presence of a mid half-bit-pair transition for zeros implies that the carrier energy per bit for a 0 is about two-thirds of that of a 1. So message 0s are more easily corrupted than 1s. This contrasts with the

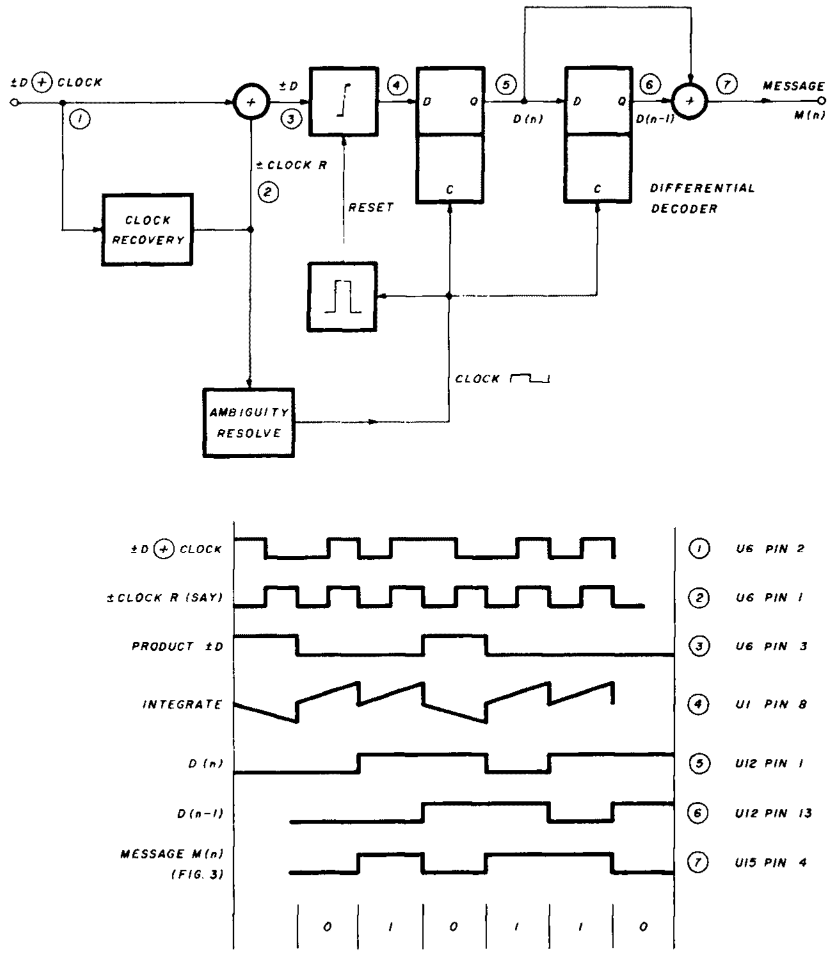


fig. 9. Message recovery: with noise present, the product ±D will be perforated and the triangles in trace 4 will become ragged. The differential decoder compares present data with the previous bit. If they are the same, the message bit is 1; if they are different, the message bit is 0. Thus the polarity of data is unimportant.

whole-bit decoder, where 0 or 1 data bit errors are equally likely but *two* message bits are always corrupted together, though less frequently.

acknowledgements

This article originally appeared in the British publication, *Electronics and Wireless World*, formerly *Wireless World* (October and November, 1984). Sincere thanks are due to editor Philip Darrington for permission to adapt the original text and illustrations for U.S. publication.

Many colleagues deserve mention, especially Trevor Stockill, G4GPQ, for encouragement, PCB layout facilities, and comparative testing with other decoders; Ron Broadbent, G3AAJ, of AMSAT-UK; Janet Miller, for letting me hog our home computer; and Cambridge Consultants Limited, for the free use of facilities.

A double-sided, legended, plated-through, printed circuit board is available from the author at the address indicated on page 50. The price, which includes shipping by air is £ 20.

references

1. M. R. Davidoff, *The Satellite Experimenter's Handbook*, American Radio Relay League, 1984. (Available from Ham Radio's Bookstore, Greenville, New Hampshire 03048, $00.00 postpaid.)
2. AMSAT-UK, London, E12 5EQ, England. Decoder alignment test tape, £ 7.50; "Oscar-10 Operating Manual", £ 5; Telemetry decoding software for BBC (Acorn) microcomputer, on cassette £ 7.50; PCB £ 20. Prices include packing and postage from UK to USA. A bank draft in sterling is requested (cash at own risk). A stamped addressed envelope must accompany ALL inquiries. AMSAT-UK depends on donations.
3. J. R. Miller, "Data Decoder for UOSAT," *Wireless World*, Vol. 89, No. 1568, May, 1983, pages 28-33.
4. A. Viterbi, *Principles of Coherent Communication*, McGraw-Hill, 1966.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

electron-hole theory exposed as fraud

Have you been bemused and confused by the electron-hole therory? Do charges, valence bonds, and the 3/2-power law make you nervous?

A startling discovery by Mark Persons invalidates all of this claptrap! In a recent issue of *Radio World*,[1] Mark reveals how electronic equipment really works:

For many years, young electronic technicians have been taught the "hole" theory of electronics. This theory explains how electrons move along conductors and semiconductors. The explanation has been good enough to satisfy or keep at bay anyone who might otherwise question the theory.

However, after a number of years working in the broadcast industry, I have come to realize the "hole" explanation may not be correct.

My theory, which has been proven time and again by personal observation, is that electronics works on smoke. *Yes, that's right. I recently learned that every manufacturer encapsulates a certain amount of smoke in every piece of electronic component he builds. The smoke is what does the work.*

You have probably noticed that a component will quit working when the smoke leaks out. I've documented this many times and it conclusively proves my theory. My theory sure beats the "hole" theory. I've never seen holes in a wire, and why don't electrons pour out of the end of the wire, if the wire is broken?

I say Mark Persons is RIGHT. I've seen smoke many times, but I've never seen an electron. Hats off to this pioneer whose discovery will be celebrated each April in the years to come!

more on VCR RFI

The subject of video cassette recorder RFI seems to keep coming up. It's a tough problem, and will probably get worse, according to Bill Pasternak, WA6ITF, who writes:

VCR-RFI is becoming a major problem and unless the manufacturers return to a higher quality product as was the case with the earlier models, I am afraid that there is little that can be done to solve the problem.

While broadcast station VCRs are designed to be immune to relatively strong rf fields, this is not true with the consumer machines. For the past five years, the manufacturers have been concentrating on reducing size and bulk and thereby the cost of manufacture by eliminating as much of the internal metallic construction as is possible. In most cases, the modern VCR consists of one or two printed circuit boards on which are mounted all of the electronic components. The only shielding is that of "tin cover plates" soldered over individual circuits that must be shielded to operate. The boards are, for the most part, secured to the plastic mainframe of the VCR and grounding between boards and chassis-of-transport is done with No. 18 wire. This construction technique, combined with the operating frequencies of the unit leaves it wide-open to interference.

The home VCR package is too small and confined to properly shield it without chancing damage of components on the PC boards. Today it appears to be "build them as cheaply as you can so the local discounter can sell them for under $300."

Unlike a TV receiver which can be effectively shielded and protected, this cannot be done with the modern home VCR. Only the equipment manufacturer can solve the problem, and as long as we go to cheaper, plastic construction, the problem will worsen rather than improve.

Finally, if you are not trained on how to service a VCR, don't even open it up to see what's inside. This is one piece of consumer electronics that should only be serviced by a highly skilled technician.

I have written a book that covers the entire spectrum of video recording from the VRX-1000 to the home VHS machine. It is titled Videocassette Recorders: Buying-Using-Maintaining *and it is available from TAB Books. If any reader of your column is contemplating the solution of VCR-RFI, I urge them to read this book, or any other*

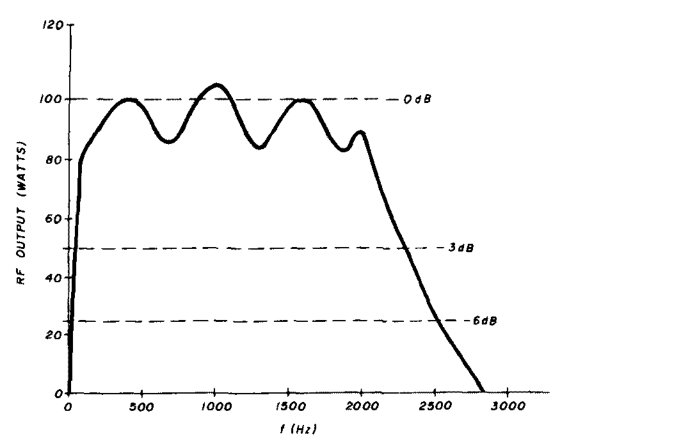


fig. 1. Representative audio response of the Yaesu FT-980 using audio generator at microphone jack and measuring RF power output. (Curve run by NE5S.)

good book on the subject, before they open up the case. Once they read it, however, they will understand why they should turn their problem over to a professional.

Hard words! But WA6ITF has been in video recording since the late 1950s, when Amperex introduced the famous VRX-1000 broadcast video recorder/reproducer. He emphasizes a serious problem that looks like it will only get worse in 1985!

transceiver frequency response

In the October,1984, issue of *ham radio* I wrote about the audio frequency response of various SSB transceivers. I received a note from Emile, N5ES, who writes:

I ran a test on my FT-980 (see **fig. 1**). *While the fluctuations don't seem too bad, it is my opinion that the whole response curve seems 300 to 400 Hz too low. Of course, there are other factors to consider, such as microphone response. I use a SHURE 444 and have gotten consistently good quality reports. Wouldn't it be nice if we had some standards in this respect!* (See "Microphone Calibration," by Daniel Peters, NY6U, *ham radio*, June, 1984, page 73.)

In this regard, Steve, K6FS, says: *The point that concerns me is the apparent confusion among Amateurs between speech "quality" and communication "intelligibility." By no stretch of the imagination can "quality" or "fidelity" be applied to the band-limited, relatively low signal-to-noise, high distortion conditions prevailing in Amateur SSB service. On the other hand, a 2800 to 3000 Hz wide passband will yield adequate intelligibility under typical Amateur conditions, if properly placed in the speech spectrum.*

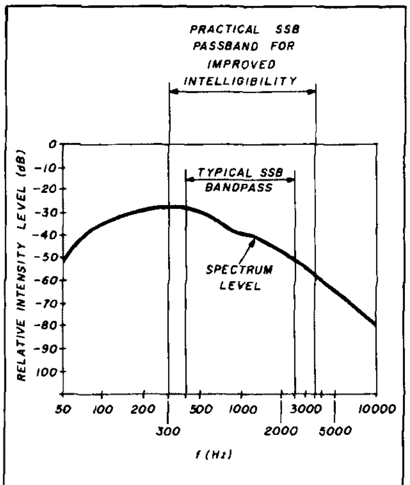


fig. 2. Typical (400-2600 Hz) and improved (300-3500 Hz) SSB passband (long-term average speech spectrum). Figure is adapted from *The Speech Chain* by Denes and Pinson, Bell Telephone Laboratories (1960). (*Data courtesy of K6FS.*)

Results of a good amount of research indicate that practical passband limits are 300 and 3000 Hz (possibly as high as 3300 Hz) with an in-band ripple of plus or minus one decibel. (See "Defining the Decibel" by Michael Gruchalla, *ham radio*, February, 1985, page 51, and "Better Sounding SSB" by Richard L. Measures, AG6K, *ham radio*, February, 1984, page 58. — Ed.)

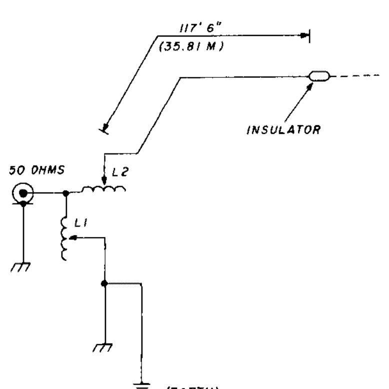
If it is necessary to narrow the transmitted passband, then intelligibility can best be protected by raising the lower limit to as high as 1000 Hz rather than shrinking the passband symmetrically. Voice frequencies below 1000 Hz are not as important to intelligibility as are those above.

One other point of confusion seems to exist: the terms for expressing filter passband limits. Usually measured in voltage terms, filter characteristics are often expressed as Hz between the "6 dB-down" points. This is equal to the half-power points "3 dB-down," in power terms. (A dB is a dB is a dB. — Ed.)

Measurements based upon power, such as shown in your article in the October, 1984, issue of ham radio *(see pages 109-111), were evidently based on output power measurements. If this is so, the distance between half-power points (shown on right-hand ordinates) delineate much narrower passbands than the "6 dB bandwidth" shown for each filter. Thus, by conventional terms, the IC-730 has a 2000 Hz passband (400-2400 Hz), the KWM-1 about 1600 Hz (550-2150 Hz) and the modified TS-830, 1800 Hz (400-2200 Hz). My suspicion is that all three would sound "muddy," cutting off the critical higher frequencies, as they appear to do.*

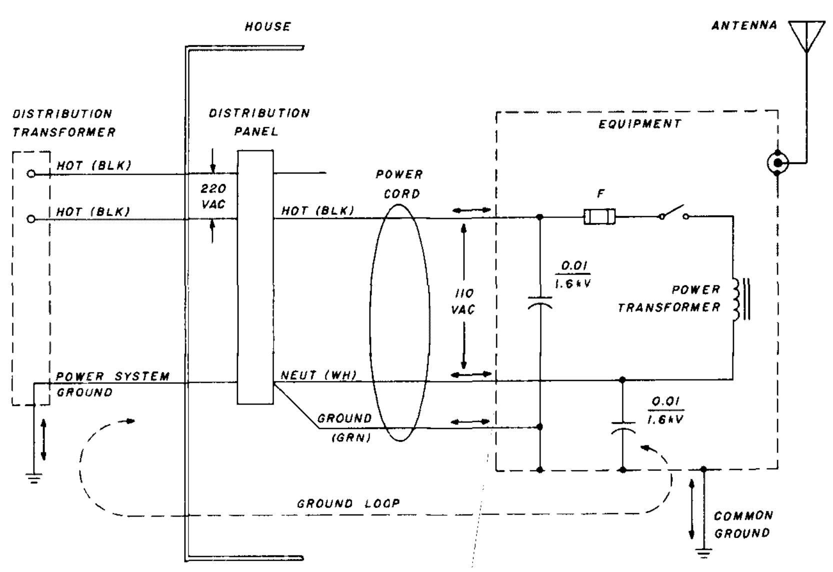
I note that my TS-130 service manual directs that carrier insertion be adjusted so as to set the −6 dB points at 400 and 2600 Hz. That's 2200 Hz bandwidth — tolerable but hardly optimal.

I am enclosing a copy of the basic spectrum level curve by Denes and Pinson, published in 1960 by Bell Telephone Laboratories. I have drawn in some passbands showing typical and suggested SSB filter character-



L1 = 10 turns, 2-inch diameter, 8 turns per inch, No. 14 wire.
L2 = 20 turns, 2-inch diameter, 8 turns per inch, No. 14 wire.
Coil L1 adjusts match. L2 resonates antenna. Mounts coils at right angles.

fig. 3. 160-meter Marconi antenna.



Note: Wrapping power cord of transmitter around a ferrite core will decouple it from equipment ground from an RF point of view.

fig. 4. RF ground currents reach earth ground via power line as well as by radio ground.

istics to illustrate the relationships discussed (*see* **fig. 2**). *Note how much energy the human voice produces below 500 Hz. It's not essential to intelligible communication.*

Well, it looks to me as if the passband filter in most Amateur SSB equipment cuts off too soon in the HF-voice region. Passing voice frequencies out to about 3500 Hz can improve intelligibility and not widen the spectrum of the signal appreciably. Most Amateur signals I've heard are wider than their voice passband anyway, mainly because of flat-topping in a linear amplifier stage (the "all-knobs-to-the-right" syndrome).

the 160-meter Marconi antenna revisited

My remarks on a 160-meter antenna in my July, 1984, column included a discussion of a practical Marconi antenna and matching network for 160 meters. After six months of use, I've come up with a better, simpler and even cheaper unit. The new design is shown in **fig. 3**.

The Marconi is matched to the 50-ohm antenna part of the transmitter through an L-network, which consists of a shunt inductor and a series capacitor. The capacitor consists of a shorter than resonance antenna — thus it costs *nothing*. The inductor (L1) is quite small and can take the form of a tapped coil. This arrangement eliminates the expensive high-capacitance variable capacitor required for the popular L-network that most Amateurs use.

The antenna is cut for the high frequency end of the band (2 MHz) and has a passband of about 75 kHz between the 2:1 SWR points. A small series inductor (L2) is added to the antenna to operate it lower in frequency.

So for the price of two inexpensive inductors, it's possible to construct a Marconi antenna that will work at any point in the 160-meter band.

a few words on ground current

The "mirror image" in the ground makes up the missing portion of the Marconi antenna, and power lost in ground resistance is subtracted from the total power. One problem with 160-meter operation is that the antenna is large with respect to the residence and the electric wiring therein, and it's easy to get unwanted coupling into the power lines that shows up as TVI and RFI in nearby entertainment equipment. (When I first went on 160 meters a few years ago, after a 40-year absence from the band, I was chagrined to find that the ceiling light in the living room lit up every time I transmitted.)

It is not easy to keep ground currents where they belong, since the transmitting equipment is connected directly to the power system ground by means of the power cord (**fig.4**). If a radio ground is added to the transmitter, two ground return circuits exist and a ground loop is formed in which high levels of current can flow. This circulating current can show up as mysterious manifestations in nearby radio and TV receivers as they, in turn, may be coupled back into the power system ground.

Since utility companies demand power system grounding, there's not much that can be done about it. The correct approach, therefore, is to isolate the transmitting equipment (from an RF point of view) from the utility ground, which is often located wavelengths away at the main distribution transformer.

The solution is to wrap the power cord of the transmitter around a ferrite core, forming an RF choke that isolates the equipment from the utility ground and makes the RF flow to ground via the radio ground attached to the equipment.

While this technique is a *must* for a Marconi antenna, it can also be useful with other antenna types. Regardless of the antenna type you have, if you have RFI problems, try wrapping the power cord of your transmitter (and amplifier, if you use one) around a ferrite rod. Wrap the power cord of the entertainment device around another ferrite rod, too. You might be surprised at how it helps clean up interference.

In my case, I taped two rods together so that the winding form was large enough for the bulky power cord. I got nine turns of line cord around the rods and then tied the power line into position at each end of the core. A recommended rod for the job is the Amidon R-33-075-1200, 12 inches long (30.48 cm) and 3/4-inch (1.90 cm) in diameter. For bands higher than 160 meters, the Amidon R-33-075-750, 7.5 inches (19 cm) long and 1/2-inch (1.27 cm) in diameter, will suffice. (The rods have a permeability of 800.) If a toroidal core is desired, the Amidon FT 240-43 can be used. It is 2.4 inches (6.1 cm) outer diameter and has a permeability of 850.

new list of EME operators

An up-to-date list of all 2-meter EME (moonbounce) operators has been compiled by Lance, WA1JXN. If you would like to have a copy, send a business-size SASE to me at Varian EIMAC, 301 Industrial Way, San Carlos, California 94070, and ask for the "EME List." Please enclose five first-class stamps, or five IRCs, for copying and postage. The list provides calls, addresses, and equipment used at active 2-meter EME stations throughout the world.

references

1. Mark Persons, "Field Service," Radio World, November 1, 1984, page 14.

ham radio

a state-of-the-art electromagnetic jargon generator

***The active field discriminator circuit** presented in this article operates on the principle of balanced product isolation. Signals from the electromagnetic vector multiplier and one parasitic signal coupler are combined with the output of an external harmonic amplitude detector. The resulting waveform is routed through the isotropic polarization generator for processing, before being applied at the output to drive an orthogonal distortion filter. (See block diagram,* **fig. 3**). *Possible applications include circular wave oscillator adjustment, as well as optimized linear frequency amplification.*

Impressive, isn't it? The above paragraph from one of my previous articles generated considerable excitement in the technical community, inspired two doctoral dissertations, and ultimately led to the Nobel Prize in Linguistic Obfuscation. But now the secret is revealed: the text above, along with all the rest of my previous technical articles, was generated by a computer. And here, for the first time in print, I reveal the secret of my literary success.

The technique upon which the state-of-the-art electromagnetic jargon generator is based was pioneered by social scientists, perfected by government employees, and has long been the mainstay of the legal profession. It involves no more than generating lists of appropriate buzzwords and catch phrases and combining them in a more or less random manner to produce a desired effect. Frequently three separate columns of words are supplied; thus, creating a ponderous technical term becomes no more complex than ordering dinner in a Chinese restaurant. Simply choose an adjective from Column A; a noun from Column B; and a noun from Column C. Add a fortune cookie ("You will meet an attractive stranger and be disappointed . . .") and a cup of hot tea, and you're ready to go. The result is the generation of phrases that sound important but mean absolutely nothing!

table 1. "Starter list" of technical terms selected by author for optimal obfuscation potential.

column A	column B	column C
linear	wave	amplifier
circular	frequency	oscillator
elliptical	phase	mixer
orthogonal	distortion	filter
isotropic	polarization	detector
harmonic	amplitude	coupler
parasitic	signal	generator
electromagnetic	vector	multiplier
balanced	product	isolator
active	field	discriminator

origin of the specious

Jim Buss, formerly K0QWI, provided the inspiration for this article. As a technical manager at the NASA Johnson Space Center in Houston, Jim generates reams of paperwork daily, including such classic phrases as: Integrated Management Options (IMO), Total Organizational Flexibility (TOF), and Systematized Policy Projection (SPP). Why not, he suggested, apply his literary technique to the fields of microwave and electronic communications?

Why not, indeed? **Table 1** contains a three-column "starter list" of words judiciously selected to meet your technical jargon requirements. Mix and match at will. By changing the suffix of the words in Column C (such as "genera*tor*" to "genera*tion*"), you can create grammatically correct terms guaranteed to fit practically anywhere in a sentence. To automate this process, I have provided, in **table 2**, a BASIC program listing designed to generate up to 1000 unique terms. How's that for Parasitic Distortion Generation?

Remember, Electromagnetic Wave Isolation requires the use of active phase detectors in combination with at least one elliptical polarization coupler to result in a harmonic vector discriminator of unparalleled quality. Now, reread all of my previous *ham radio* articles[1-19] and see how many of these terms you recognize!

By H. Paul Shuch, N6TX, 14908 Sandy Lane, San Jose, California 95124

table 2. Microsoft™ BASIC program facilities generation of up to 1000 incomprehensible technical terms.*

```
10 '---------> JARGON.BAS <--------------------
20 '            Rev. A, 13 Aug '84
30 '            by N6TX
40 '  COPYRIGHT (C) 1984 MICROCOMM
50 '
60 '  Generates totally meaningless combinations
70 '  of Microwave/Electronics buzzwords!
80 '
90 '-----------------------------------------
100 CLR$ = CHR$(26) ' Defines Clear-Screen String
110 PRINT CLR$
120 PRINT "DO YOU WISH OUTPUT ROUTED TO:"
130 PRINT
140 PRINT "                     PRINTER (P)"
150 INPUT "            or     SCREEN  (S)";PR$
160 IF PR$="P" OR PR$="p" OR PR$="S" OR PR$="s" GOTO 200
170 PRINT CLR$
180 PRINT "YOU MUST RESPOND WITH 'P' OR 'S'" : PRINT
190 GOTO 120
200 '-----------------------------------------
210 PRINT CLR$
220 '            Random Number Seed entered here
230 PRINT "'JARGON.BAS' generates random combinations of"
240 PRINT "Microwave/Electronics buzzwords, for inclusion"
250 PRINT "in technical manuscripts.
260 PRINT
270 PRINT "To start the randomization process, it will be"
280 PRINT "necessary to enter a Seed Number."
290 PRINT
300 INPUT "ENTER ANY NUMBER HERE:  ",S
310 RANDOMIZE (S)
400 '----------------------------------------------------
410 PRINT CLR$
420 INPUT "How many technical terms do you wish to generate";N
430 IF N>0 GOTO 460
440 PRINT : PRINT "number entered must be greater than 1."
450 GOTO 420
460 IF N = INT(N) GOTO 490
470 PRINT : PRINT "number entered must be an integer."
480 GOTO 420
490 PRINT CLR$
500 '----------------------------------------------------
510 '        ARRAY LISTED HERE
520 DIM A$(10,3)
530 A$(0,0) = "LINEAR": A$(0,1) = "WAVE": A$(0,2) = "AMPLIFIER"
540 A$(1,0) = "CIRCULAR": A$(1,1) = "FREQUENCY": A$(1,2) = "OSCILLATOR"
550 A$(2,0) = "ELLIPTICAL": A$(2,1) = "PHASE": A$(2,2) = "MIXER"
560 A$(3,0) = "ORTHOGINAL": A$(3,1) = "DISTORTION": A$(3,2) = "FILTER"
570 A$(4,0) = "ISOTROPIC": A$(4,1) = "POLARIZATION": A$(4,2) = "DETECTOR"
580 A$(5,0) = "HARMONIC": A$(5,1) = "AMPLITUDE": A$(5,2) = "COUPLER"
590 A$(6,0) = "PARASITIC": A$(6,1) = "SIGNAL": A$(6,2) = "GENERATOR"
600 A$(7,0) = "ELECTROMAGNETIC": A$(7,1) = "VECTOR": A$(7,2) = "MULTIPLIER"
610 A$(8,0) = "BALANCED": A$(8,1) = "PRODUCT": A$(8,2) = "ISOLATOR"
620 A$(9,0) = "ACTIVE": A$(9,1) = "FIELD": A$(9,2) = "DISCRIMINATOR"
630 '------------------------------------------------------------------
640 '            PRINT HEADER
650 IF PR$ = "S" OR PR$ = "s" THEN 700
660 LPRINT "      ELECTROMAGNETIC JARGON BY MICROCOMM"
670 LPRINT "      --------------- ------ -- ---------"
680 LPRINT
700 '------------------------------------------------------------------
710 '            START LOOP HERE
715 PRINT CLR$
720 FOR I = 1 TO N
730 '            GENERATE RANDOM 3-DIGIT NUMBER
740 X = INT (RND * 1000)
750 A = INT (X / 100)
760 B = INT (X / 10) - (10 * A)
770 C = X - (100 * A) - (10 * B)
780 PRINT A$(A,0);TAB(17);A$(B,1);TAB(34);A$(C,2)
1000 '-----------------------------------------
1010 IF PR$ = "S" OR PR$ = "s" THEN 1030
1020 LPRINT A$(A,0);TAB(20);A$(B,1);TAB(40);A$(C,2)
1030 '-----------------------------------------
1040 NEXT I
1050 PRINT : PRINT
1060 IF PR$ = "S" OR PR$ = "s" THEN 1100
1070 LPRINT : LPRINT
1080 LPRINT : LPRINT
1090 '---------------------------------------------
1100 INPUT "TYPE <return> TO CONTINUE, 'Q' TO QUIT ",D$
1110 IF D$ = "Q" OR D$ = "q" THEN GOTO 1130
1120 GOTO 630
1130 END
```

*This program is also available for the Apple IIE. Send SASE.

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ham radio

the ham notebook

improved carrier suppression for the MC1496

The MC1496 has been around for many years and has enjoyed widespread popularity in many balanced modulator design applications. Recently, while attempting to build a 50-MHz DSB generator using this device, I came across several small circuit improvements that will increase the carrier suppression levels, especially at the higher frequencies.

The first improvement involves the use of a bifilar wound toroidal tank circuit that takes advantage of the inherent self-balance of the bifilar windings, yielding a noticeable improvement in carrier feedthrough. In addition, by feeding the signal into the center tap of the bifilar winding through a series choke arrangement, the windings are isolated from unbalanced ground effects.

Normally carrier balance is set through a DC biasing adjustment using a trim pot. The addition of two small-value trimmer capacitors from each side of the MC1496 output pins to ground substantially improves carrier suppression by allowing further balance of the RF tank circuit. If interaction occurs, some readjustments may be necessary for optimum results.

Finally, while the recommended carrier injection level is 60 mV RMS, I found that by varying the LO drive slightly above or below this level *often* offered improved carrier suppression levels. Any variation in drive level, at 50 MHz, will upset the circuit balance and require further carrier balance readjustments.

While these changes apply to the MC1496 at 50 MHz, one may wish to try similar modifications, at lower frequencies, to improve the expected level of performance from this device. It is likely that these techniques could be applied to the SN76514 and SL6440 IC mixer devices.

Peter Bertini, K1ZJH

bulkhead connector

The type 83 bulkhead connector (83-1F) is useful as a panel-mounted coax feedthrough; however, these fittings are expensive and are available to most hams only through mail order. The PL258 (83-1J) double-female connector is inexpensive and is carried by Radio Shack stores, but it has no provision for panel mounting and attempts to solder a flange to it invariably result in melted polystyrene dielectric. A weekend project can be thwarted by the lack of a suitable fitting.

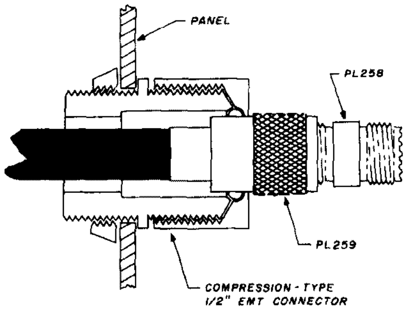


fig. 1. Cross-section of homemade bulkhead connector.

A satisfactory, if not aesthetically perfect substitute for the 83-1F will provide mechanical stability and can be assembled from locally-available parts. The outside diameter of the coupling ring of a PL259 (83-1SP) is nearly the same as that of 1/2″ EMT electrical conduit. Half-inch EMT fittings are available at most hardware stores. A compression-type 1/2-inch EMT junction box connector will grip a PL259/PL258 pair to form a sturdy panel-mount coax connector, as shown in cross-section in **fig. 1**. It is only necessary to *remember* to slide the EMT connector onto the cable *before* installing the PL259.

Gary Myers, K9CZB

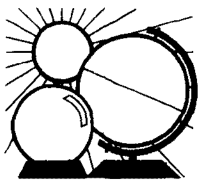
Midway Amateur Radio Club assumes management of North American TRN

The Midway Amateur Radio Club of Kearney, Nebraska, now sponsors the North American Teleconference Radio Net (TRN). TRN links together over 150 gateway stations (mostly VHF/UHF repeaters) across the US and Canada to present high quality technical and informational programs of interest to Radio Amateurs. Past speakers on TRN have included Vic Clark, W4KFC, and Senator Barry Goldwater, K7UGA.

The idea for TRN began with Ed Piller, W2KPQ, and Charlie Kosman, WB2NQV. In the early 1980's Ed and Charlie began linking repeaters by telephone to provide technical presentations as a joint project of the Long Island Mobile Amateur Radio Corps (LIMARC) and the Long Island Chapter of IEEE. However, with the telephone bridging equipment available to them, it was difficult to provide high quality audio to and from all participating repeaters. In late 1982 Rick Whiting, WØTN, a telecommunications engineer, became net manager. Rick made arrangements with Lou Appel, KØIUQ, of Darome, Inc., to use Darome's sophisticated multipoint teleconference bridges to provide the "land line" links for repeaters. The result was superb audio quality and a rapid growth in the number and distribution of gateway stations in the net. Lou will continue to be the bridge engineer in TRNs under the new net manager.

Requests for TRN information should be sent to TRN Manager, c/o Midway Amateur Radio Club, P.O. Box 1231, Kearney, Nebraska 68847-1231. (SASE please, Canada excepted.)

ham radio



DX FORECASTER

Garth Stonehocker, K0RYW

sunspot cycle views

In the October, 1984, *DX Forecaster* we discussed the present 10.7-year sunspot cycle. How accurate were our six-month predictions?

The solar flux dropped even lower than the minimum forecast for August as a result of an almost non-existent 27-day cycle variation (i.e., a nearly spotless sun) from mid-September to mid-November. October had the year's lowest recorded monthly solar flux — 74. The minimum daily value of flux so far in sunspot cycle 21 was 69 on September 29. Since October, both flux and activity have increased, approaching the February-March annual maximum of approximately 90 flux units (36 SSN).

Expect the flux to decrease toward an annual minimum during the months of July, August, and September. Expect the *daily* flux to drop down near the previously recorded low of 67 (August 25, 1954) during the summer 1986 or 1987. The sunspot cycle decline has definitely changed during 1984, from the steep decline experienced during 1982 and 1983 at a rate of 4.5 flux units per month to a leisurely rate of about 5 per *year*.

What does this mean in terms of working DX over the next few months? As the solar flux decreases, the MUF can also be expected to decrease. Although this might suggest a pessimistic view of summertime DX, the opposite is often the case: F_2 layer propagation is poorer when the F_2 layer MUFs is low; however, two compensating propagation effects also occur. The first is due to the greater number of hours of daylight in the summer, which means MUF rises earlier in the day. Also, the MUF remains higher until sunset than it does in winter. This effect is mainly felt on paths in east-west and northern directions in our hemisphere. On southern bearings, which are usually transequatorial (TE) one-long-hop in winter, the MUFs in the evening are usually lower in the summertime — i.e., not much TE propagation is available; the high electron density areas ±20 degrees from the magnetic equator just don't build up in the summertime as they do in the winter.

The other compensating propagation factor is sporadic E, which provides short-skip conditions out to 1200 miles (2000 km), with multiple hops possible. This propagation follows the sun across the sky with maximum effect at local noon for higher band DXing and near sunrise and sunset for the lower frequency bands. More detailed information on using E_s propagation will appear in next month's column.

last-minute forecast

The higher HF bands, 10-30 MHz, are expected to be very good during the first two weeks of April, with the 27-day solar flux maximum the main determining factor. Transequatorial openings should be good the second and third weeks of the month, corresponding to disturbed geomagnetic field conditions, and during the equinoctial period. The lower frequency bands, 2-10 MHz, are expected to be best during the third and fourth weeks, at least between weather storm fronts moving by your location. Look for unusual DX on east-west paths that touch the auroral latitudes, (60 to 70 degrees north latitude) during disturbed periods in the middle of the month.

The perigee of the moon's orbit (for moonbounce DX) is on the 5th, with the moon showing full phase on the 5th. There will be a short meteor shower, the Lyrid, on April 20-22, with a rate of five per hour — hardly much help for meteor-scatter DX. But a bigger shower, the Aquarid, starts before the end of April, peaks on May 5, and ends in mid-May. Its rate is 10 to 30 per hour.

band-by-band summary

Ten meters will be open to the south and southeast for a short period before local noon, to the south at noon, and to the southwest in the afternoon. The openings will be longer when the solar flux is at its 27-day cycle maximum. Even better transequatorial one-long-hop conditions will occur during disturbed periods. Listen to WWV at 18 minutes after the hour and note the geomagnetic field status announcement (A and K indices).

Fifteen and twenty meters, almost always open to some part of the world, will be the main daytime DX bands. Twenty should stay open on long southern paths into the night, while 15 will drop out in the late afternoon. Operate 15 first and move down to 20 meters. Contacts out to 5000 to 7000 miles (8000 to 11,200 km) are possible on these bands and one-long-hop transequatorial propagation may also occur, as it does on 10 meters.

Thirty and forty meters are both day and night bands. Intermediate distances 1000 to 1500 miles (1500 to 2200 km) in any direction, considered daytime DX, are better now than in SSN maximum years. Nighttime DX on these bands may be expected to offer greater distance paths than on 80 meters and, like 80, follow the darkness path across the sky. Reduced midday signal strengths and distances may occur on days of high solar-flux values, with 30-meter openings disappearing in the pre-dawn hours on the morning after the high radio-flux values occur.

Eighty and one-sixty meters will exhibit short skip conditions during the daylight hours and lengthen at dusk. These bands follow the darkness paths, opening to the east just before your sunset, swinging more to the south near midnight, and ending up in the Pacific areas during the hour or so before dawn. The 160-meter band opens later and ends earlier.

Coastal stations and those with good low-angle radiating systems will usually have the edge for working rare DX. QRN will be as low on some nights as that experienced during the wintertime DX season.

APRIL

WESTERN USA

GMT	PST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	4:00	20	20	15	10	15	10	10	20
0100	5:00	20	20	15	10	15	10	10	20
0200	6:00	20	20	20	10	15	10	10	20
0300	7:00	20	20	20	10	20	10	10	20
0400	8:00	20	20	20	10	20	15	10	20
0500	9:00	20	20	20	10	20	15	10	20
0600	10:00	20	20	20	15	20	15	15	20
0700	11:00	20	30	20	15	20	20	15	20
0800	12:00	20	30	20	15	30	20	15	20
0900	1:00	*20*	30	20	15	*30*	*20*	*20*	*30*
1000	2:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1100	3:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1200	4:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1300	5:00	*30*	20	15	*20*	*30*	*20*	*20*	*30*
1400	6:00	*30*	20	15	*20*	*20*	*20*	*20*	*30*
1500	7:00	*30*	20	15	*20*	*20*	20	*20*	*30*
1600	8:00	*30*	20	15	20	20	15	20	*30*
1700	9:00	30	20	15	20	20	15	20	20
1800	10:00	30	20	10	20	20	15	15	20
1900	11:00	30	20	10	15	20	15	15	20
2000	12:00	30	20	10	15	20	15	15	20
2100	1:00	20	20	10	15	15	10	15	20
2200	2:00	20	20	15	15	15	10	15	20
2300	3:00	20	20	15	15	15	10	10	20

MID USA

GMT	MST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CST
0000	5:00	20	20	15	15	20*	10	10	20	6:00
0100	6:00	30	20	15	15	20*	10	10	20	7:00
0200	7:00	30	20	20	20	20	10	10	20	8:00
0300	8:00	*30*	20	20	20	20	15	10	20	9:00
0400	9:00	*30*	20	20	20	20	15	15	20	10:00
0500	10:00	*30*	20	20	20	20	15	15	20	11:00
0600	11:00	*30*	20	20	20	20	15	15	20	12:00
0700	12:00	*30*	30	20	20	20	20	15	20	1:00
0800	1:00	*30*	30	20	*20*	*30*	20	20	20	2:00
0900	2:00	*20*	30	*20*	*20*	*30*	20	*20*	*30*	3:00
1000	3:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*	4:00
1100	4:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*	5:00
1200	5:00	20	*30*	15	15	*30*	*20*	*20*	*30*	6:00
1300	6:00	20	20	15	15	*30*	*20*	*20*	*30*	7:00
1400	7:00	20	20	15	15	*30*	*20*	*20*	*30*	8:00
1500	8:00	20	20	15	15	20	*20*	*20*	*30*	9:00
1600	9:00	20	20	15	10	20	15	20	*30*	10:00
1700	10:00	20	20	10	10	20	15	20	20	11:00
1800	11:00	20	20	10	10	20	15	15	20	12:00
1900	12:00	20	20	10	10	20	15	15	20	1:00
2000	1:00	20	20	10	10	20	10	15	20	2:00
2100	2:00	20	20	10	10	20	10	15	20	3:00
2200	3:00	20	20	15	15	20*	10	15	20	4:00
2300	4:00	20	20	15	15	20*	10	10	20	5:00

EASTERN USA

GMT	EST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	7:00	20	20	15	15	20*	10	10	20
0100	8:00	30	20	15	15	20	10	10	20
0200	9:00	30	20	20	20	20	10	10	20
0300	10:00	*30*	20	20	20	20	15	10	20
0400	11:00	*30*	20	20	20	20	15	15	20
0500	12:00	*30*	30	20	20	20	15	15	20
0600	1:00	*30*	30	20	20	20	20	15	20
0700	2:00	*30*	30	20	20	*30*	20	15	20
0800	3:00	*20*	*30*	20	*20*	*30*	20	20	20
0900	4:00	*20*	*30*	*20*	*20*	*30*	20	20	*30*
1000	5:00	20	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1100	6:00	20	*30*	*20*	15	*30*	*20*	*20*	*30*
1200	7:00	20	*30*	15	15	*30*	*20*	*20*	*30*
1300	8:00	20	20	15	15	*30*	*20*	*20*	*30*
1400	9:00	20	20	15	15	*20*	*20*	*20*	*30*
1500	10:00	20	20	15	15	20	*20*	*20*	*30*
1600	11:00	20	20	15	10	20	15	20	*30*
1700	12:00	20	20	10	10	20	15	20	20
1800	1:00	20	20	10	10	20	15	15	20
1900	2:00	20	20	10	10	20	15	15	20
2000	3:00	20	20	10	10	20	10	15	20
2100	4:00	20	20	10	10	20	10	15	20
2200	5:00	20	20	15	15	20*	10	15	20
2300	6:00	20	20	15	15	20*	10	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

the weekender

a DC dummy load

Anyone who works with heavy-duty batteries or with low-voltage DC power supplies develops a keen appreciation for any handy way to test them under load. The instrument described in this article can be built inexpensively in a single evening. It can put the heaviest duty Amateur power supply to the test and reveal a great deal about it.

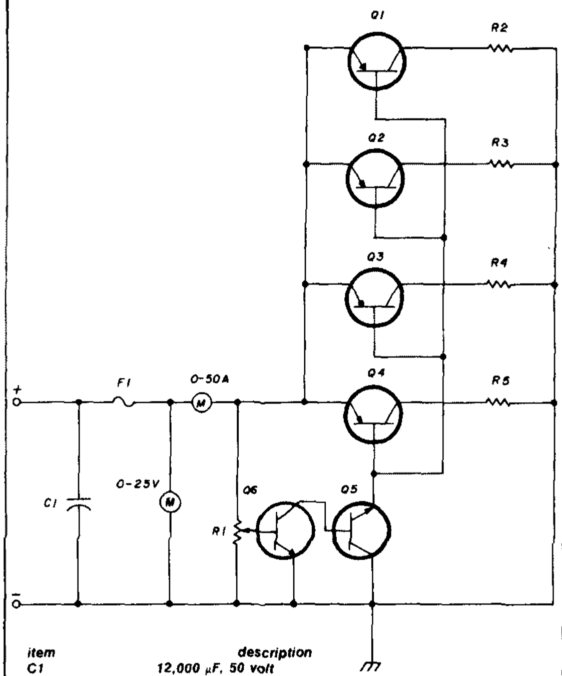
This instrument monitors load current and voltage while allowing smooth adjustment of the load from zero to maximum. With the addition of optional features, it can test transformers for current-handling capability, test the strength of automotive or stationary batteries, and check voltage regulation and overall performance of most low-voltage power systems. Better yet, you can build it from junk box parts or from Radio Shack parts costing less than $25. You'll need some power transistors to absorb energy, an ammeter, a voltmeter, a control potentiometer, a couple of heat sinks, and other assorted hardware.

design requirements are flexible

The circuit for the DC load tester I built is shown in **fig. 1**. The idea for the unit was not original with me; I dimly remembered seeing something like it in an old magazine,[1] but I couldn't find the article right away, so I started from scratch.

Because I have some heavy-duty power sources around, including a 50-ampere, 14-volt power supply and a big bank of lead-acid batteries, I wanted a load that could take a lot of current and push these sources hard. With an early breadboard model, I was mystified when a pair of 2N3055 transistors failed well before they reached their rated maximum collector current of 15 amperes each. Then it dawned on me: these transistors have a maximum dissipation rating of 115 watts each, for a total of 230 watts. Their efficiency, of course, is zero, so all that power is dissipated as heat. I was asking those two transistors to dissipate 15 volts at 20 amperes — a total of 300 watts.

By George L. Thurston III, W4MLE, 2116 Gibbs Drive, Tallahassee, Florida 32303



item	description
C1	12,000 μF, 50 volt
F	fuse — value depends on design
Q1,Q2,Q3,Q4	Radio Shack 2043 (PNP)
also suitable:	ECG 219, ECG 180, ECG 61, ECG 184, ECG 68, ECG 34, ECG 93
Q5	RS 2019
Q6	2N2222
R1	1,000 ohms, 1 watt
R2,R3,R4,R5	0.05 ohm, 10 watt

fig. 1. AC/DC load tester.

With a little figuring, I decided I could live with a unit that would dissipate 600 watts, so I chose somewhat higher power PNP transistors, Radio Shack 2043s, rated to handle 15 amperes and 150 watts dissipation for $2.19 each.

I mounted four 2043s on large, black anodized heat sinks. The collector resistors, which serve to keep the current evenly divided among the four transistors, had to be made up from 5-watt, 0.1-ohm units bought from a local supply house at about 45 cents each. Suitable resistors can often be found at flea markets for as little as ten cents or so. The actual value of the equalizing resistors is not critical, as long as you keep them below about 0.15 ohm, but they should be the same for each of your paralleled transistors.

Any number of different types of transistors would work equally well as long as all transistors in any one project are alike and combined power dissipating ability is sufficiently large. To economize on enclosure size and on the number of heat sinks and equalizing resistors, use the fewest high-wattage, high-current transistors that will do the job. NPN or PNP units are equally suitable, though the bias control circuit must be different for each type and, of course, the emitter-

collector connections must be reversed (see **fig. 2**).

In the breadboard models a 25-watt, 50-ohm wire-wound rheostat and some fixed resistors were used to control bias on the transistor and consequently the amount of current they drew from the source. However, the wire-wound resistor did not provide smooth control, and a better method was needed. A smaller transistor could be used to regulate the base current and could itself be adjusted with an ordinary potentiometer. But what kind of base-current control transistor would work? Obviously, the control transistor would have to handle the total base current of the combined load transistors, which would be asked to deliver a maximum of 60 amperes.

The Radio Shack 2043s I used had a current-gain ratio (h_{FE} or Beta) of about 20. (The Betas of individual transistors differ quite widely sometimes, even within the same production batch.) This meant that if I wanted 60 amperes from the load transistors, I would need 60/20, or 3 amperes of base drive current.

My junk box yielded a Radio Shack No. 2019, a transistor rated to handle 10 amperes of collector current and with an h_{FE} of about 20. To deliver 3 amperes, it must have about 0.15 amperes (150 mA) of base drive. This is easily obtained from an wire-wound potentiometer hung across the input source. Even less base current would be required by a transistor with a higher Beta (or h_{FE}). The TIP 120, which can deliver 5 amperes of collector current, has a Beta of about 1,000.*

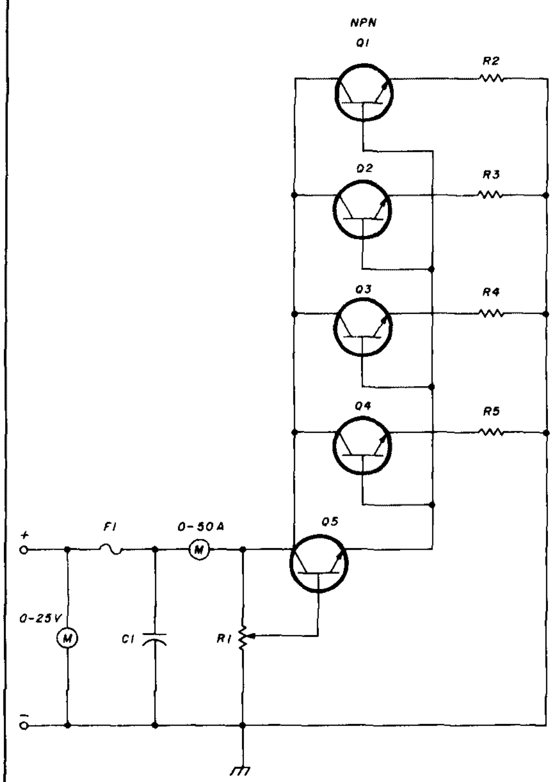


fig. 2. NPN transistors require the emitter-collector connections to be reversed.

metering

I used a commercial 30-ampere meter with an external shunt. (**Figure** 3 shows how to use a sensitive micro- or milliamp meter to measure current in several ranges.) The meter measures the voltage drop developed across one of the equalizing resistors. By proper selection of values for R6 and R7, you can choose any convenient current range. One range should go slightly above the maximum current for which the unit was designed.

If you plan a maximum of, say, 30 amperes, the high-range meter should read about 50 amperes full-scale. Assuming two load transistors, half the total current will flow through each resistor. If each resistor is 0.1 ohm, the drop will be 1.5 volts.

If you use a meter with 100 microampere full-scale sensitivity, the calibration potentiometer, R6, will need to be at least 1.5 volts divided by 0.0001 amperes, or 15,000 ohms. Even though the meter's internal resistance will be approximately 1000 ohms, the potentiometer will compensate for it. Use any convenient value between about 15,000 and 25,000. A little circuit board type potentiometer will work fine.

*The final model used an ordinary 1/2-watt carbon pot in the base of a 2N2222, which drives the base of the RS 2019 driver transistor.

To calibrate, use an ammeter of known accuracy in the input lead to the load tester and adjust the calibration potentiometer until the meter readings agree. If your meter has a 0-5 scale, you can read it as 0-50 amperes. A reading of 1 would indicate 10 amperes.

Another way to calibrate, if you don't have a high-range ammeter for reference, is to put a VTVM or FET-VOM across one of the equalizing resistors and increase the load current until you reach some predetermined voltage. Then, knowing the *value of the* resistor, you can calculate the current flowing through it and set your calibration potentiometer accordingly. In the example above, a reading of 1.5 volts would correspond to a total current of 30 amperes.

For convenience in reading light loads, a second potentiometer can be used to provide a 0-5 ampere range and a panel switch can select the desired range.

testing

Aside from meter calibration and wiring errors, there isn't much to test for except oscillations. Transistors, while capable of amplifying, will sometimes oscillate independently. To prevent this, select transistors with low Beta and low maximum frequency ratings.

If oscillations occur, they can be detected with a scope or with an RF probe on a high-impedance voltmeter. The solution is usually a matter of bypassing something; adding a 0.1 to 0.47μF metal film, solid tantalum or ceramic capacitor, from the transistor base to ground with the shortest possible leads, will generally solve the problem. In multiple-transistor circuits such as our AC/DC load tester, it may be necessary to bypass several transistor bases. If oscillations persist, try bypassing emitters, collectors — anything above ground potential. A last resort might be to insert very small value resistors in series with one or more base leads. In the circuit used, however, the transistor Betas are quite low and oscillations were no problem.

If you use bypass capacitors, be sure their voltage ratings are high enough to withstand the highest voltages you're likely to apply to the input terminals. And remember that when you're testing an AC source through the bridge rectifier, you will encounter voltages about 1.4 times higher than the AC RMS voltage.

applications

The most obvious use of this device is to see which goes up in smoke first — the load or the power supply. But it is capable of much more sophistication than that.

Monitor both the output voltage and current. Then plot a graph of the relationship to get a good record of the quality of regulation of your supply.

Check the fold-back current limiter that's built into many supplies. At what value of load current does it go into action? It may shut down too soon, depriving you of some output capability of the supply. Or it may not shut down soon enough, subjecting your supply to unnecessary stress. Set the load to draw whatever amount of current you think is safe for the supply, then adjust the shut-down threshold to the point at which it turns off the supply.

Monitor the output of the supply with a scope while you slowly increase the load. How much current can you draw before ripple appears in the output? That may tell you something about the design of the filter and the regulation of the rectifier output.

The scope will also tell you if the supply will tend to oscillate at certain load settings — a possible cause of poor regulation and regulator burnout, not to mention TVI or birdies in nearby broadcast radios.

As any auto mechanic knows, the health of a lead-acid battery is best tested under load. A fully-charged battery whose voltage sags to 10 or 11 volts under a load of 10 to 15 amperes is sick! Voltmeter readings across individual cells or hydrometer readings on each cell will spot the defective one and confirm your diagnosis.

Using a bridge rectifier at the input, you can check

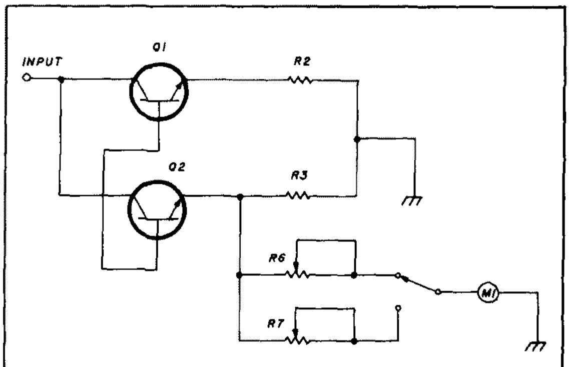


fig. 3. External resistors enable sensitive meter to be used on several ranges. (See text for component values.)

the performance of a transformer before you build it into a piece of gear. The unregulated output of any transformer, when rectified and filtered, will drop under load. Plot measurements of voltage and load current on graph paper to check performance of the filter. The voltage will fall gradually as load increases until the transformer is delivering all it is designed to provide. After that, the voltage will begin to fall faster with each additional ampere of load current because of copper losses, core saturation, hysteresis, and other problems that crop up when the transformer is overloaded. That point should appear on your graph as a slight "knee" in your curve.

rectifier bridge

To add a bridge rectifier to your dummy load, all that's necessary is to provide two additional terminals for the AC input and to hook them to a bridge (see fig. 4). The positive output of the bridge is connected directly to the positive input terminal of the dummy load. The negative bridge terminal goes to the negative load terminal. The bridge output can be left connected to the DC input terminals. It won't conduct with positive voltage applied to the positive load terminal.

Radio Shack offers a 25-ampere, 50-volt rectifier bridge for less than $3 that will work well in this application, provided you never demand more than 25 amperes or apply more than 50 volts to it. For really heavy-duty applications, you may want to build a bridge from discrete diodes, each rated at 30 amperes or more at 10 volts or more.

With AC applied to the bridge, it will handle twice the maximum rated current of the individual diodes, since two diodes are working at once on alternate halves of the cycle. Thus, a bridge with 35 ampere diodes would safely deliver 70 amperes of current — more than adequate for testing most Amateur-service transformers.

The bridge input could be hooked directly to the DC input of the dummy load and would provide automatic

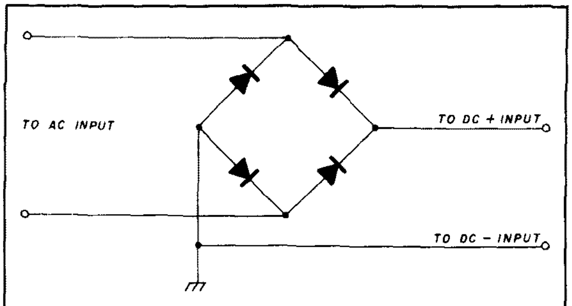


fig. 4. Addition of bridge rectifier adds to versatility of load tester.

polarity correction. No matter how the DC input is connected to the AC bridge terminals, the output from the bridge will always be the same. But there's a good reason for not hooking up the bridge this way. When you put DC on the AC terminals of the bridge, two diodes in series work simultaneously, just as with AC. But they work *all* the time — not alternately, as in a 50 percent duty cycle. That means that DC input to the bridge must be limited to the current each diode will handle, or about half the AC rating. Thus, the Radio Shack bridge would be good for only 12.5 amperes when DC is applied to its input terminals.

another refinement

You will probably find that by advancing the load control potentiometer on your dummy load you can increase the load current enough to disintegrate either the load, the power supply, or both. You can protect against this in several ways:

- Install a fold-back current limiter with an adjustable threshold.
- Use a fuse low enough to blow if you exceed a safe current.
- Limit current by putting a maximum-load resistor in series with the transistors.
- Limit current by padding the control potentiometer with a fixed resistor.

The fold-back limiter may be unnecessarily fussy and complex. The fuse may not blow until the transistor junctions have gone to glory.

Although W7RXV uses fuses as equalizing resistors in his dummy load design, a hazard is involved besides their slowness compared to the junction.[1] Fuses are seldom exactly the same. If one blows before the other, much of the load will be shunted to the other transistors, overloading them and blowing their junctions before the fuses go.

A maximum-load limiting resistor is feasible. A 0.5-ohm resistor inserted between the transistors and ground will prevent the load impedance from going below that value. Thus, with 20 volts applied, the load current would be limited to 40 amperes, even if the transistors shorted or were turned fully on. The same resistor would limit current to 30 amperes at 15 volts and to 20 amperes at 10 volts. This would considerably reduce the range of the load or the effectiveness of the resistor. Additionally, the resistor would have to be rated at 800 watts to handle 40 amperes since $P = I^2R$. More realistically, it would have to be rated at about 450 watts to handle currents up to about 30 amperes. Such resistors are bulky and difficult to find.

The most satisfactory choice could be a resistor in series with the control potentiometer. For NPN load transistors, it would be inserted between the positive input terminal and the potentiometer. For PNP transistors, it would be inserted between the potentiometer and ground. Even this is not foolproof, because it will still be possible to exceed the dissipation rating or the current rating of the load transistors under some conditions.

reference

1. Evert Fruitman, W7RXV, "The Smoke Tester," *73*, October, 1976, page 159.

bibliography

Roos, John, K6IQL, "The Power Waster," *73*, January, 1981, page 108.

ham radio

controlled vertical radiation rhombics

part 2: antenna erection and performance

Four towers, careful siting and plenty of wire yield topnotch results

Very few projects can be completed without compromise, and erecting a rhombic antenna is no exception. In this project, the compromise lay between tower height and low-frequency operation. An antenna height of about 65 feet was the maximum desired; this meant that 7 MHz would be the lowest design frequency. The desire to have no more than eight wavelengths on the 28-MHz band established the leg length. As pointed out in part 1 of this article, an eight-wavelength array produced an 8-degree beamwidth, and anything less was not desired. Thus two wavelengths on 7 MHz equals 277 feet. The antenna performs well on the 7, 10, 14, 18, 21, 24, and 28 MHz bands and to a lesser extent on the 1.8- and 4-MHz bands.

Based on the height and leg length, the tilt angle for the various bands was determined, with special emphasis on the lowest and highest bands — 7 and 28 MHz, respectively. These bands would determine the greatest width between the side towers and the greatest length between the end towers. **Figure 1** shows the horizontal layout of the designed rhombic, emphasizing the 7 and 28 MHz configurations. Note that a 2 A angle (see part **1**, **fig. 3**) of 50 and 20 degrees are required for the 7 and 28 MHz configurations, respectively. This requires the let-out of the rhombic from the side towers of about 72 feet to go from 7 to 28 MHz (see table 1). The resulting takeup at the end tower is about 45 feet.

The determination of these parameters is a matter of trigonometry and is not detailed here.

geographical bearing and antenna layout

The determination of the location of the four towers is the crucial point of the design; once the towers are set in concrete, they can't be moved. Two points to consider in this regard are the desired bearing of the rhombic from true north and the allowance for a gap between pulleys and down lead. The downlead supports a counterweight, which must clear the tower as the counterweight is raised and lowered. An additional three feet between pulley and tower is advisable.

Unlike a Yagi, whose beamwidth may be anywhere from 40 to 60 degrees, the eight-wavelength rhombic will have a beamwidth of only 8 degrees. Therefore, accuracy in determining both the true bearing to the desired reception area and the physical positioning of the four towers supporting this antenna is especially important. Check all measurements and calculations carefully. (See references 1 and 2 for how to determine bearings.)

To determine the bearing of my antenna, and to set the ground posts properly, I borrowed a transit and stood where the fixed-antenna tower was to be erected. I sighted Polaris — the North Star — while KA4ECM, my wife Millie, stood about 400 feet away, shining a flashlight toward me. I lowered the transit, to a point parallel to the ground still keeping it pointed squarely in the direction of Polaris, while KA4ECM walked slowly in an east-west direction. As soon as I spotted her light in the transit viewfinder, I signaled for her to stop. She then planted a ground post at that point.

The following day we reset the transit to align on that ground post, and swung it 46.8 degrees from North, inserting a second post at the distant point. (Accuracy of ±1 degree is recommended.)

By **Henry G. Elwell, Jr., N4UH**, Route 2, Box 20G, Cleveland, North Carolina 27013

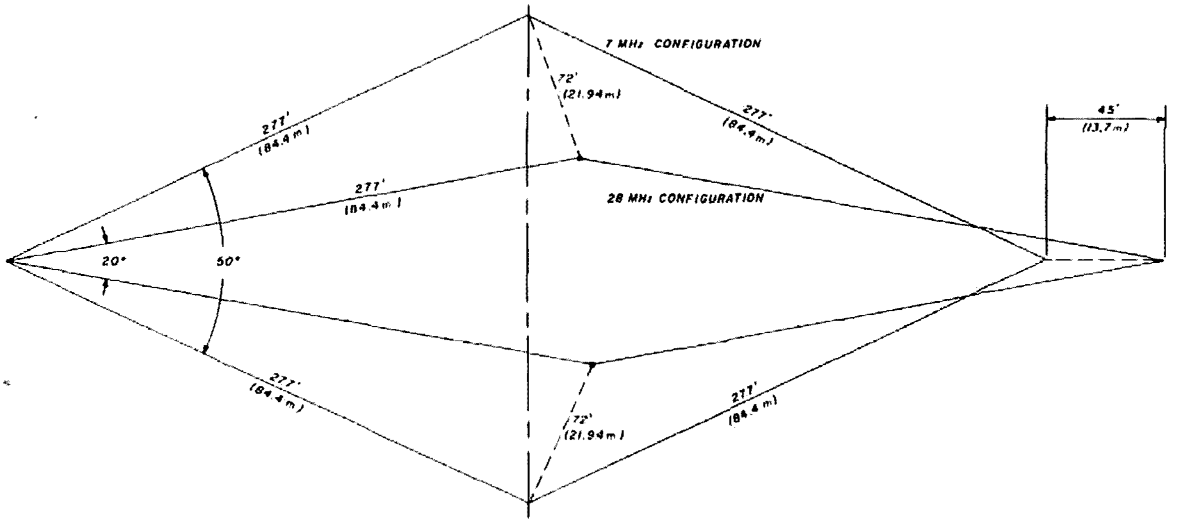
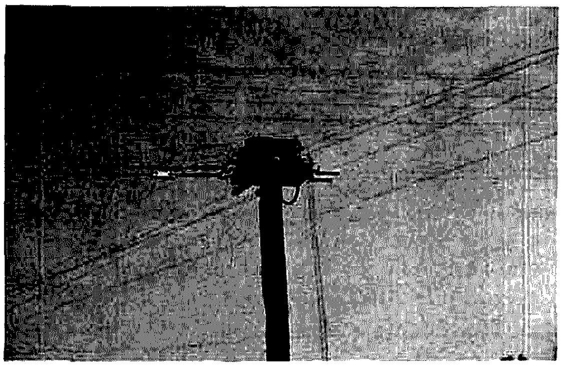


fig. 1. Horizontal layout of controlled vertical radiation rhombic at the lowest frequency (7 MHz) and the highest design frequency (28 MHz band).



Reversing switch at center of rhombic field.

table 1. Correlation between amount of "let-out" and antenna height.

side tower let-out, feet	antenna height feet
0	65.00
10	61.75
20	59.00
30	57.00
40	55.00
50	53.75
60	53.50
70	52.25

We stretched a length of nylon string between the two posts, each representing an end tower, and then took accurate measurements (using a steel tape) to (a) the point representing the center line of the side towers, and (b) the point representing the end point of the 28-MHz rhombic, corrected for the increased distance needed for pulley-tower separation. (Plan on installing a pulley on the fixed end so that the whole array can be dropped to the ground when necessary.)

We then placed the transit on the nylon string at the point representing the side tower's centerline. Ninety-degree right and left bearings were made and posts temporarily set at distances of approximately 120 feet. We strung nylon cord between these posts, took accurate measurements along the cord, and then drove stakes representing the two side towers into the ground.

Even though Polaris is easy to spot with the naked eye, it may be difficult to locate with the transit or telescope. Because of its great distance from Earth, its light reaches the telescope in parallel rays, making magnification difficult. Taking your bearings on a clear, windless — and not too cold — night minimizes the discomfort of an already difficult task.

The terrain over which my rhombic had to be erected was generally level but fell off quite sharply to the east. Various tower heights were necessary to produce an antenna that would be parallel to sea level. Two were 70 feet; a side tower had to be 80 feet; and the far end tower was 100 feet. The necessary tower heights were determined from topographical maps.

If the rhombic is to be erected on ground with a uniform slope extending for at least 1000 yards in front

table 2. Control of the vertical angle of radiation is by paying out the side tower cables.

let-out distance (feet)	angle 2A (degrees)	vertical angle of radiation — degrees 7.2 MHz	10.1 MHz	14.2 MHz	21.3 MHz	28.6 MHz
0	50.0	26	16.0	5	beam splits	
10	45.8	28	20.0	11	beam splits	
20	41.6	30	22.5	16	beam splits	
30	37.6	31	24.0	18	6.0	beam splits
40	33.4	32	26.0	21	10.0	7
50	29.2	33	27.0	22	14.0	11
60	25.0	34	27.5	23	16.0	13
70	21.0	35	28.0	24	17.5	15

of the antenna, the practice is to make the front and rear poles approximately equal in height in order to bring the major axis of the antenna parallel to the average ground slope.

The erection of towers has been well covered in the literature. Substantial towers, well guyed in three directions according to the manufacturer's specifications, are necessary. One point to remember is that one guy set should be directly behind the vector force of the antenna at each tower.

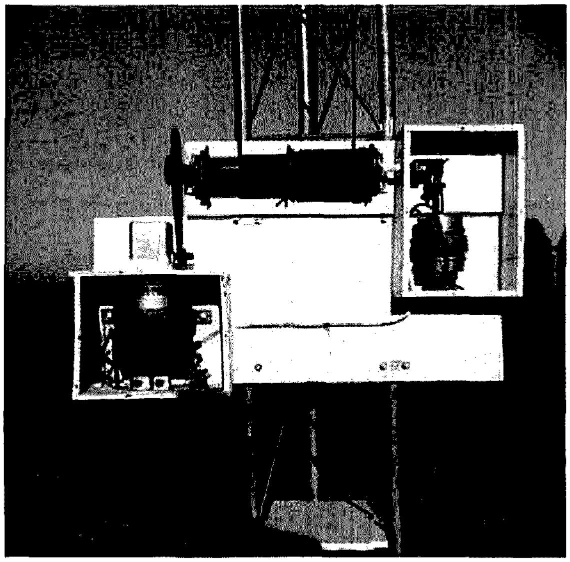
In private communications, Marshall Etter, W2ER, who was chief engineer at the old RCA overseas receiving station in Riverhead, Long Island, stated that tests showed it was not necessary to interrupt the guy wires with insulators for rhombic arrays. However, if the towers are also to be used for mounting Yagis, the use of guy insulators should be considered.

As to choice of antenna wire, the military specifies high strength, 40 percent conductance, using three strands of No. 12 AWG copperweld wire, with a rated breaking point of 2433 pounds. Other wire, such as No. 6 AWG (0.162 inch), 40 percent conductance copperweld, may also be used. W2ER stated that seven-strand No. 16 AWG bronze wire was used in the RCA Rocky Point installations. He advises against the use of solid wire, which tends to vibrate in long spans. I used a special stranded steel core, wrapped with copper, used for aircraft trailing-wire antennas.*

For maximum strength, treat the rhombic as four long-wire antennas. Terminate each leg at an insulator. At the side tower, terminate both the left and right legs on the same insulator holes, connecting the two legs with a flexible jumper soldered to each leg for a good electrical circuit (see fig. 2). The flexible jumper provides slack when changing the tilt angle for different bands or propagation conditions.

The end tower antenna legs are terminated in insulators, and two insulators connected to a pear ring

*Marshall Etter, W2ER, has a limited supply of wire, insulators, and other rhombic antenna construction materials. Inquiries (enclose SASE) should be addressed to W2ER at 16 Fairline Drive, East Quogue, New York 11942.



Side tower drive system. Motor drive is at left; take-up spindle, center; synchro position transmitter, right. Note the chain drive from motor to spindle to provide increased torque.

(see fig. 3). The pear ring provides a strong, easy way of connecting the three forces: two antenna legs, and the opposing force of the restraining cable.

standing wave ratio

Once the complete system was operational, SWR was measured. Figure 4 shows the average overall SWR from 160 meters through 10 meters. The increase in SWR in the 160-meter band may be expected because the exponential lines were designed for a minimum frequency of 3.5 MHz and the 4:1 balun cannot be expected to operate properly over such a wide frequency range. Although the big rise in SWR in the 21-MHz band is not understood, it is believed to be associated in some way with the balun. Operation in the 15-meter band is excellent, and with all open wire lines, losses due to a 3:1 SWR are minimal.

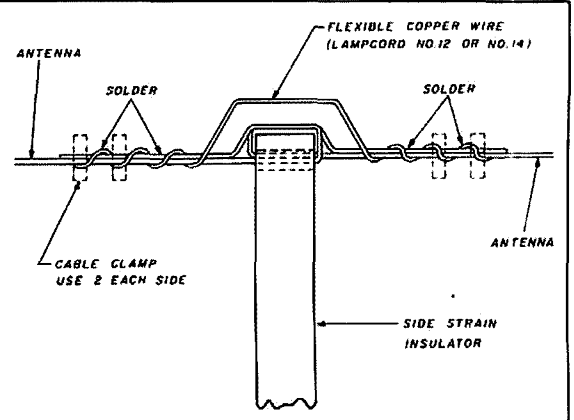
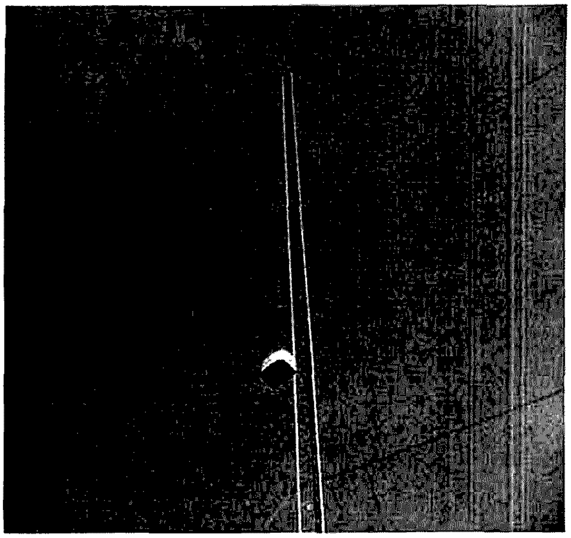


fig. 2. Connection of antenna leg at side towers to obtain antenna strength and good electrical connection between antenna legs.



End tower; 220-pound weight is shown in 40-meter position.

I have long believed in using tubes, which are very forgiving on SWR excursions, in the output amplifier stages. Use of solid-state outputs would require investigation and correction for SWRs greater than 2:1 or suffer the reduction in output levels which such transceivers automatically provide.

report on results

It will take many more months of operation and evaluation to explore fully all the capabilities of the CVR rhombic. After almost a year's operation during its development period, results can be reported in qualitative terms. Additional testing with data on the effects of changing the configuration for operation on a specific band as well as other details may be the subject of another article.

The rhombic has been used on all current Amateur bands from 160 through the 10-meter band. Foster graphs[3] determined the expected results to be as shown in **table 2**. A report on its actual performance, band by band, follows.

- **160 meters.** The rhombic has a constant 3:1 SWR over the entire band. Using an FT102, phone contacts were made from this QTH to the Virgin Islands, to Canada, and west to Texas. Definite front-to-back response was noticed when reversing direction of fire. Comparisons were made against an 80-foot W2LL vertical with apparent advantages going to the vertical for longer distances, and to the rhombic for shorter distances.

- **80 meters.** The rhombic, 1 wavelength on a leg, has an SWR less than 2.5:1 over the entire band. It exhibits definite gain over a W2LL vertical (60-foot tower with TH7DX Tribander on top) for DX. Friends in the New York City area said the rhombic's signal was stronger than that of any other antenna I've ever used. One of the first contacts was with VK6HD, long path at 2157Z. There is about a 15-dB front-to-back difference when reversing antenna fire direction. The beamwidth is noticeable, ZL4PO/C, within the beamwidth, gave me a 58 when all other East Coast stations were getting 56 to 57. However, Australian stations, which are outside the beamwidth by 23 degrees, give me reports comparable to other East Coast stations.

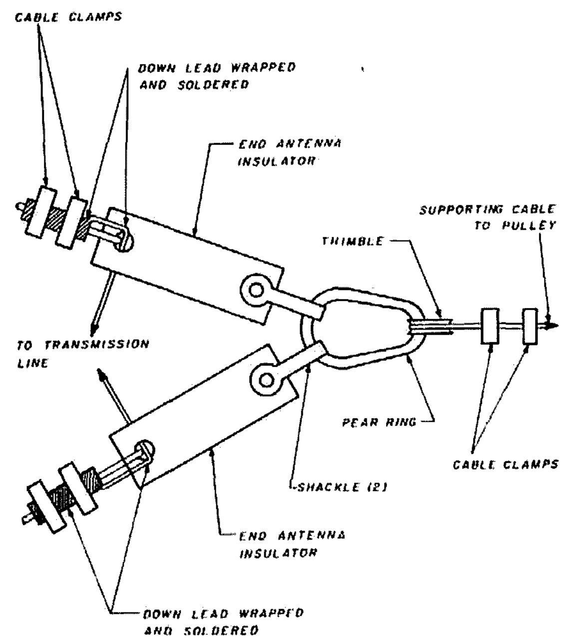


fig. 3. Pear ring used for terminating antenna and supporting cables.

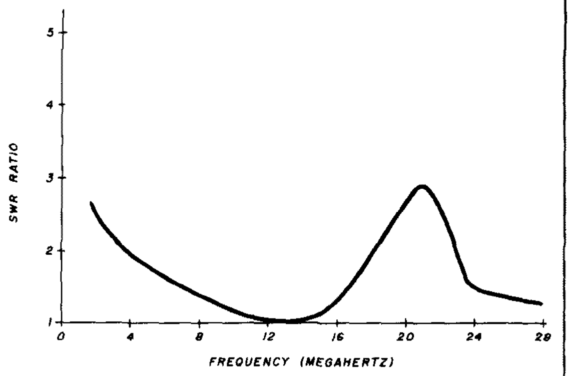


fig. 4. Graph illustrates *smoothed* SWR ratio versus frequency.

• **40 meters.** The rhombic exhibits a sharp change in SWR over this band: 3:1 from 7.0 MHz, dropping to 1.5:1 at 7.2 MHz and remaining there over the rest of the band. These SWR changes are peculiar to my particular setup. Being terminated, the SWR should be reasonably flat over the entire band. The rhombic is 2 wavelengths on a leg on this band. Reversing the direction of fire results in a 30 dB change in signal strength.

I've used the rhombic on this band mainly in contests, with phone results being the most informative. Because contest signal reports are meaningless, antenna effectiveness has been judged by the fact that, except for a single exception, my station received the first reply when a European stood by for Stateside calls on a stated frequency. No other 40-meter antenna was available during this evaluation period to supply comparative reports. However, the "feel" is that the rhombic, in its favored directions, is equal to or better than 3- or 4-element Yagis.

• **30 meters.** The rhombic, 3 wavelengths on a leg, has an SWR of approximately 1.5:1 across the band, and a front-to-back ratio of about 30 dB. Its signal really shines on this band because most of the antennas in competition with the rhombic are relatively simple. In the 40-meter configuration (on the 30-meter band) the rhombic's vertical angle of radiation (VAR) is at its lowest angle — 16 degrees — so ground reinforcement on this band (30-meter) has not been of major significance.

• **20 meters.** The rhombic, 4 wavelengths on a leg, has an SWR practically flat over the entire band; that is, 1.2:1 or less. Operation is with the side insulators 30 feet from the side towers for open-band operation into Europe where its vertical angle of radiation is about 18 degrees. Experience has shown that for band opening, long-path operation, and band closing operation, a 40-meter configuration with a VAR of 5 degrees on 20 meters provides very impressive signals. My first CQ on this band with the FT102 during band-opened conditions was a sufficient reward for all the effort put into siting and construction of the rhombic.

The comparison antenna on this band is a Hy-Gain TH7DX at 60 feet; this is a tribander beam on a 24-foot boom with two driven elements, a director, and a reflector. It is an excellent antenna that puts me among the top callers in most pileups.

The results of comparisons with European stations show a 1-1/2 to 2 S-unit advantage of the rhombic over the TH7. Front-to-back ratio is about 25 dB. The narrow beamwidth of the rhombic is noticeable on this band; thus, signals into Australia are superior on the TH7, as expected, because Australia is off the 3-dB edge of the main lobe by almost 30 degrees.

• **16 meters.** No operation has been accomplished on this band, although a quick SWR check indicated an SWR of 1.7:1.

• **15 meters.** The rhombic is 6 wavelengths on a leg on this band and for some unknown reason, not investigated, the antenna system has an SWR ranging from 3:1 to 2.2:1 over the band.

Preliminary controlled vertical radiation operation was accomplished over the entire configuration change of the rhombic, — that is, with side insulators at zero feet to 70 feet. Because it takes about 30 seconds for a 10-foot increment change of the side insulators, I thought it best to take qualitative measurements on a broadcast station. Radio Berlin at 21.6 MHz was used for the test. A 25-dB change in signal strength was noted between zero feet and 40-50 feet, with a buildup of 5 dB from that point to 60 feet and 70 feet. Much more work has to be done in this area.

• **12 meters.** Again, no operation is permitted on this band, although a quick SWR check indicated a 1.2:1 SWR ratio.

• **10 meters.** On this band the rhombic is 8 wavelengths on a leg and its operation is truly awesome to someone like me, whose biggest 10-meter antenna was a 5-element Yagi at 55 feet. I first operated on this band in February, 1983, with the rhombic in a 40-meter configuration. At that time I had not yet fully explored the beamwidth of the rhombic. But I foolishly asked Roger, N4ZC, with his large antenna farm to work VKs and ZLs with me to see how our signals compared. In both countries his 6-element Yagi at 90 feet was 2 S-units better than my rhombic.

Since that time I've realized that even ZL is too far beyond the 8-degree beamwidth of the rhombic on 10 meters, and the 40-meter configuration simply does not operate properly on 10 meters.

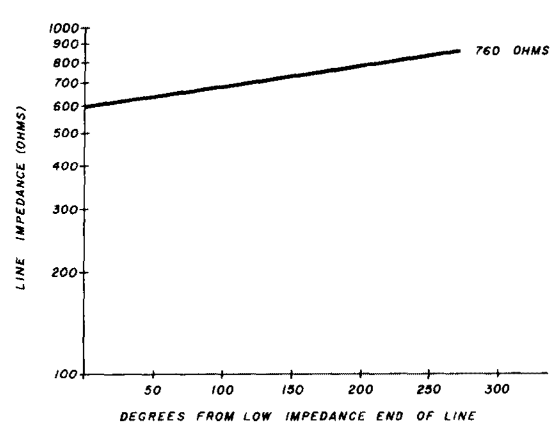


fig. A1. An exponential transmission line is used to transform impedances.

Ten meters opened again in the Fall after I had gained more experience with the rhombic and adjusted it to operate on this band. In the RSGB 10/15 meter contest in October, although not noted for my speedy contest operation, I worked 61 UK stations in 25 minutes because of the strength of the signal I was putting into Europe; "First W heard this morning," and "Loudest signal on the band," were pleasant to hear.

The general opinion appeared to be that the rhombic was 3-4 S-units stronger than the TH7. However, the narrow beamwidth is very noticeable on this band. With 45 ARRL countries within the beamwidth of the rhombic, the population density of Europe was being saturated, which was the original objective of the project.

acknowledgements

A project of this size cannot be accomplished without help and I wish to thank W2ER, WD4KJZ, WD4FFX, W2IRC, W2KXD, Alan Sielke, K1AA, W2LL, and my wife Millie, KA4ECM, whose contributions made it all possible.

references

1. Jerry Hall, K1PLP, "Bearing and Distance Calculations by Sleight of Hand," *QST*, August, 1973, page 24.

2. Chester H. Brent, WB4GVE "Aim Your Beam Right," *73*, June, 1976, page 122.

3. Donald Foster, "Radiation of FM Rhombic Antennas," *Proceedings of the I.R.E.*, Volume 25, October, 1937, page 1327.

appendix

determination of an exponential line

Exponential transmission lines are useful in transforming impedances. The values at any point along the line can be determined using graphical or mathematical approach.[A1,A2] Since these references may not be available to everyone, a description of their method follows.

graphical method

This method uses a minimum transmission-line dimension of one-half wavelength drawn on semi-log paper.

- Mark the vertical scale with the desired impedance range. All cases will probably be from 100 to 1000 ohms.
- Mark the horizontal scale in wavelengths or electrical degrees.
- Mark a point at the 0 degree location with the desired input impedance.
- Mark a point corresponding to the desired output impedance and degrees point. The separation between these two points should be at least one-half wavelength (180 degrees).
- Draw a line between the two impedance points.
- The required characteristic impedance can now be read from the graph at all intermediate points along the line.
- Determine the necessary line configuration from each impedance point from the formula:

$$a = P\ 10^{Z/276} \text{ 2-wire line} \qquad \textbf{(A1)}$$

$$a = \frac{P}{\sqrt{2}}\ 10^{Z/138} \text{ 4-wire line side-connected} \qquad \textbf{(A2)}$$

$$a = \sqrt{2}\ P\ 10^{Z/138} \text{ 4-wire line, cross-connected} \qquad \textbf{(A3)}$$

where a = distance between wires of transmission line, in inches
P = radius of wire, in inches
Z = desired line impedance

Table A1 lists wire radius for various size wires. **Table A2** shows impedance variation along an exponential transmission line as a function of location and line spacing.

Example: Design an exponential 2-wire line to go from 760 ohms to 600 ohms using No. 14 bare copper wire, 3/4 wavelength (270 degrees). **Figure A1** is first drawn and then a table set up to show distance from the 600-ohm point, the impedance representing that distance, and (from the above equations) the line spacing required; see **table A2**. For greater accuracy, use the second method.

mathematical method

The impedance, Z, at any point on an exponential transmission line can be mathematically described as:

$$Z = Z_s\, e^{2s\theta} \qquad \textbf{(A4)}$$

where Z_s = the input or sending end impedance in ohms
s = line length in wavelengths (1 wavelength = 360 degrees)
θ = is a transformation function

The transformation function can be determined from the desired characteristics of the transmission line; i.e., input impedance, Z_s, and desired output impedance, Z, at the end of the line of s wavelengths. Its equation is:

$$\theta = \frac{1}{2s}\ ln\ \frac{Z}{Z_s} \qquad \textbf{(A5)}$$

To solve, determine:

- input and output impedance; Z_s and Z.
- line length; (minimum of 1/2-wavelength at lowest operating frequency); s.
- solve for θ.
- solve for Z for each selected value of s; let s be no greater than 20 degrees; 10 degrees preferably.
- determine the necessary line configuration for each value of Z using line spacing formulas shown in **graphical method**.

Example: Design an exponential 2-wire line to go from 600 ohms to 760 ohms using No. 14 bare copper wire, 3/4-wavelength long (270 degrees).

table A1. Radius of common wire gauges.

wire size	radius (inches)
8	0.0642
10	0.0509
12	0.0406
14	0.0320
16	0.0254
18	0.0200

table A2. Impedance variations along an exponential transmission as a function of location and line spacing.

distance from 600-ohm point (degrees)	characteristic impedance (ohms)	line spacing (inches)
0	600	4.78
20	611	5.23
40	620	5.64
—	—	—
250	750	16.69
270	760	18.14

table A3. Mathematical and graphical method results compared.

distance from 600-ohm point (degrees)	characteristic impedance (ohms)	line spacing inches
0	600.0	4.78
20	610.6	5.22
40	621.4	5.71
—	—	—
250	747.2	16.31
270	760.0	18.14

Steps 1 and 2 have already been determined. It is then necessary to solve for θ, which will become a constant for this example; Z_s is also a constant.

$$\theta = \frac{1}{2 \cdot 0.75} \ln \frac{760}{600} = 0.158$$

Then solve for Z at 10 or 20 degree intervals and tabulate. Twenty-degree intervals are shown in **table A3.**

$$Z = 600\, e^{(2 \cdot \frac{20}{360} \cdot 0.158)} = 610.6$$

Line spacing is then determined:

$$d = 0.032\,(10^{600/276}) = 4.78 \text{ inches}$$

It will be seen that a slight discrepancy exists between the values of Z obtained from the graphical method and from the mathematical method because of difficulty in reading the curve.

For a given spacing, a four-wire line will give a much lower impedance than a two-wire line, and a cross-connected four-wire line will give an even lower impedance (than the same dimensioned side-connected four-wire line). Also the larger the wire-diameter wire, the lower the impedance for a given line separation. These factors may be used to arrive at your design of an open wire exponential line.

references

1. Edmund Laporte, *Radio Antenna Engineering*, Chapter 3, figure 3.81 and Chapter 4, page 422.
2. John D. Ryder, *Networks and Fields*, Prentice-Hall Inc., Chapter 6, Section 6-13.

ham radio

run RTTY
on your Timex

Get on RTTY — inexpensively

Before any computer can be used for RTTY, several problems must be solved. This article shows how these problems can be overcome in adapting this popular low-cost computer for use on this interesting mode. Although the hardware described is configured specifically for the Timex-Sinclair, the principles involved are applicable to all types of home computers.

basic requirements

A terminal unit (TU) decodes the Mark and Space tones from a receiver. An input/output (I/O) port enables the decoded signals to communicate with the computer and software tells the hardware when and how to perform.

A simple serial I/O port and a TU are described. Assembler routines are included for initializing, reading from, and writing to the port.*

RTTY is transmitted in Baudot code. Characters consist of a start bit, five data bits, and stop bits. ASCII characters contain seven data bits and an eighth bit sometimes used as a parity check bit. Each bit in each character must be checked or generated by the computer.

Communication to and from the Timex/Sinclair is most easily accomplished via the cassette port, which represents only one bit of a parallel I/O port. The computer is synchronized with the data by writing precisely timed delays, called software timing loops, into the program.

UART aids data handling

The low price of the Timex/Sinclair computer has no doubt contributed significantly to its mass appeal. One cost-cutting measure in its design was the elimination of the video controller chip. The video display is generated by the microprocessor. When operating in the continuous display mode, the microprocessor devotes most of its time to creating the display with execution of the program carried out only *between* video frames. Consequently there is no time for the microprocessor to use software timing loops to process data at reasonable baud rates.

This problem can be solved by the addition of a serial I/O port. In a serial port, individual bit timing is handled by a Universal Asynchronous Receiver Transmitter (UART). The workload on the computer microprocessor is greatly reduced as data bits are assembled into complete characters during receive and characters are converted to serial bits during transmit by the UART.

I first considered using the popular AY series UART, but these devices must be furnished with a baud rate clock. A different clock frequency is required for each desired baud rate. Generation of a stable baud rate clock for the various baud rates used for RTTY and ASCII created yet another problem.

One suggested solution involved using an INS 8250 chip, a combination UART and baud rate generator.

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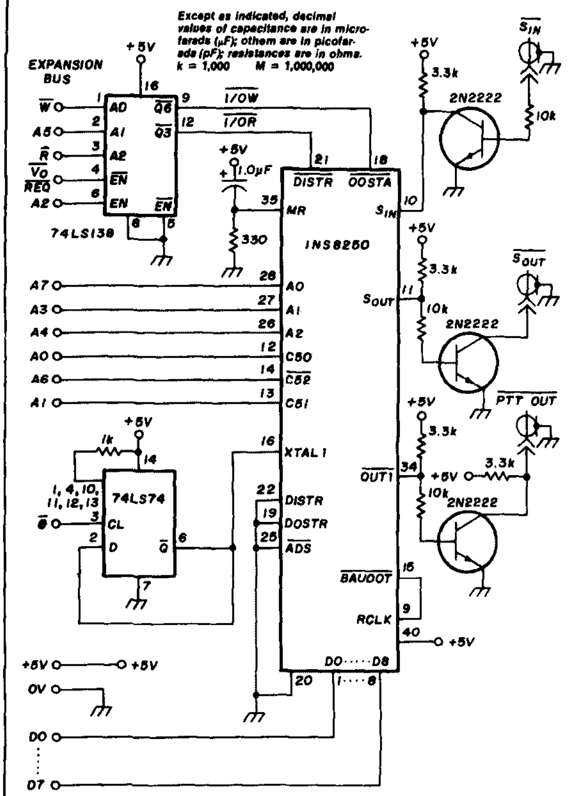
*Circuit boards for the Terminal Unit and the I/O port, as well as a full-featured RTTY/ASCII transceive program for Timex/Sinclair 1000 and 1500 computers are available from the author.



capacitors	
quantity	description
1	1.0 µF
2	0.047 µF
ICs	
1	74LS74
1	74LS138
1	INS8250
resistors	
1	330 ohm
1	1.0 kilohm
4	3.3 kilohm
2	10 kilohm
sockets	
1	14 pin
1	16 pin
1	40 pin
transistors	
3	2N2222

fig. 1. I/O port schematic.

When supplied to the INS 8250, a clock frequency of up to 3.1 MHz can be divided in frequency by any two-byte word from software to provide the baud rate clock. To implement the port it's necessary only to provide address decoding and to divide the 3.25 MHz computer clock by two.

Although the cost of the INS 8250 is about twice that of a simple UART, this is more than justified by its overall circuit simplicity: a serial port with a crystal controlled, keyboard adjustable baud rate can be built from a design using only three chips.

configuration of INS 8250

The INS 8250 is enabled by bringing pins CS0 and CS1 high and pin CS2 low. Eight internal registers are addressed by pins A0, A1, and A3. Data is strobed into or out of these registers by pins DOSTR and DISTR.

table 1. Initialization and Read/Write routines. (Routines written in Z80 mnemonics.)

```
;               ROUTINE TO READ A CHARACTER FROM PORT
;
;   ROUTINE RETURNS WITH NEW CHARACTER IN 'A' REGISTER IF CHARACTER
;   READY, ELSE RETURNS WITH 'A' REGISTER = FFH
;
;
RXCHR:  IN      A,(LNSTAT)      ;READ LINE STATUS REG
        BIT     0,A             ;IS A CHAR READY?  NZ = YES
        JR      Z,NORDY         ;GO IF NOT CHAR READY
        IN      A,(OBUFF)       ;READ CHAR FROM REC BUFFER
        RET                     ;RETURN WITH CHAR IN A REG
NORDY:  LD      A,(0FFH)        ;PUT FFH IN A REG
        RET                     ;RET WITH A = FFH IF NOT CHAR READY
;
;
;
;
;               ROUTINE TO SET PORT FOR TRANSNIT
;
XMIT:   LD      A,RTTYT         ;GET MODE CONTROL WORD IN A REG
        OUT     (LNCON),A       ;ENTER WORD IN LINE CONTROL REG
        LD      A,4             ;PREPARE TO SET BIT 2 IN NODEN CONT REG
        OUT     (MCONR),A       ;ACTIVATE PRESS TO TALK OUTPUT
        RET                     ;DONE: EXIT
;
;
;
;             ROUTINE TO WRITE A CHARACTER TO THE PORT
;
;   ENTER ROUTINE WITH HL REGISTER POINTING TO CHAR TO BE TRANSMITTED
;   ROUTINE CHECKS TO SEE IF THE INS8250 IS READY TO SEND ANOTHER
;   CHARACTER AND RETURNS WITHOUT TAKING ANY ACTION IF PORT NOT READY
;   CHARACTER IS SENT IF PORT READY, AND BIT 7 OF CHARACTER IN BUFFER
;   IS SET INDICATING TO CALLING PROGRAM THAT CHARACTER HAS BEEN SENT
;
;
TXCH:   IN      A,(LNSTAT)      ;READ THE LINE STATUS REGISTER
        BIT     5,A             ;IS THE PORT READY FOR A NEW CHAR NZ = YES
        RET     Z               ;RETURN IF PORT NOT READY
        LD      A,(HL)          ;GET NEW CHAR IN A REG
        OUT     (OBUFF),A       ;WRITE CHAR TO THE PORT
        SET     7,(HL)          ;IN TIMEX/SINCLAIR THIS SETS INVERSE VIDEO
        RET                     ;DONE: EXIT
;
;
```

The Line Control register sets the number of data bits, stop bits, and various error checking options. This register also allows access to the Divisor latches for setting the baud clock, which runs at 16 times the actual baud rate.

The Line Status register is read during transmit to determine when the chip is ready for the next data byte to be sent to the Transmit register. It is read during receive to determine when a character is ready to be read from the Receive buffer. Various errors are also shown in this register.

The Modem Control register determines the output of four Modem Output pins. These could be used for such functions as Press-To-Talk, Keying, CW ID, or CW keying. (My program used only the Press-To-Talk function.) The status of four Modem Input pins can be read from the Modem Status register. One of these could be used to detect a decoded CW Mark (key down) signal.

active filter improves operation

The second piece of hardware required is the Terminal Unit. My first circuit used an XR 2211 phase locked loop demodulator. Receiver output was fed directly to the demodulator. It was immediately apparent that filtering of the signal before the demodulator would be necessary for satisfactory operation under normal QRM and QRN conditions. An

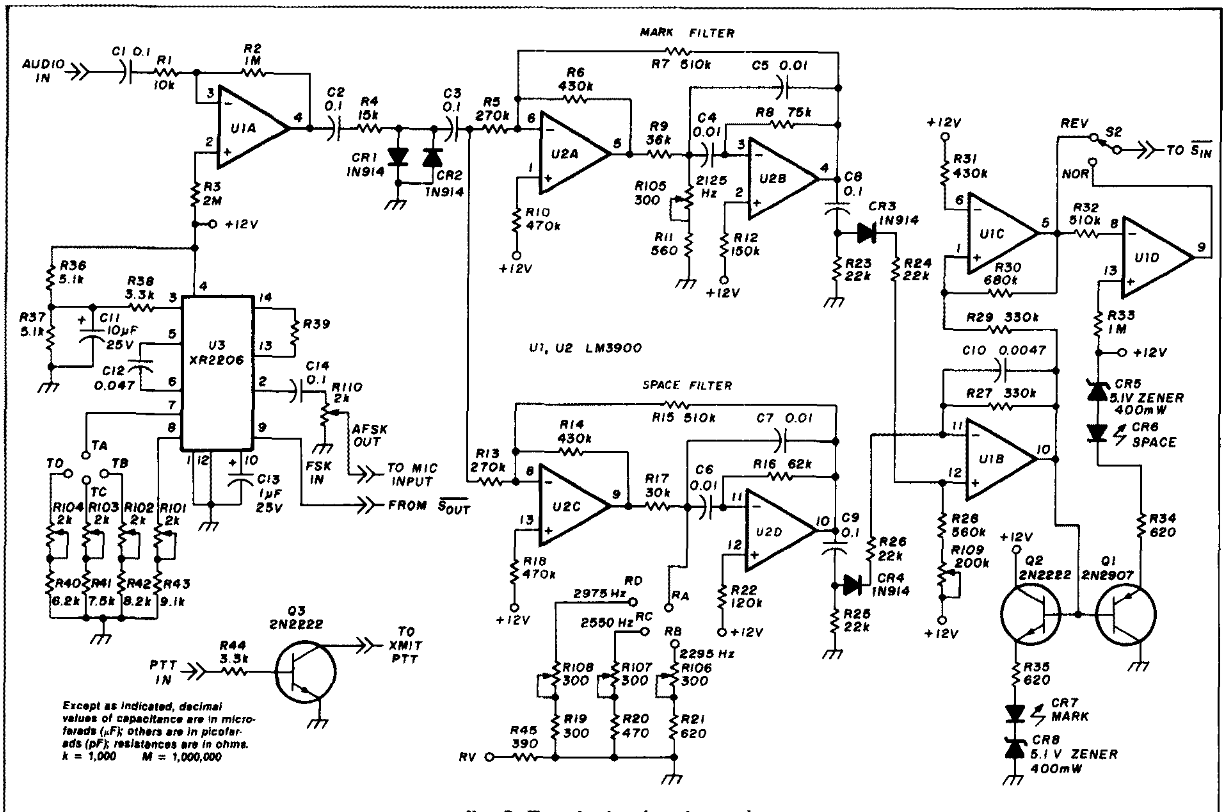


fig. 2. Terminal unit schematic.

active filter circuit for 170 Hz shift was no problem, but the circuitry became complex when provisions were added for copying signals at 425 Hz and 850 Hz shifts as well.

The design shown uses an active filter-type demodulator. It represents a reasonable compromise between performance and circuit complexity. A familiar XR 2206 circuit is used to generate the transmit AFSK tones.

operation at the I/O port

The schematic for the I/O port is shown in **fig. 1**. Power for the unit and signals from the computer are taken from the computer expansion bus. Charging of the 1.0 μF capacitor connected to pin 35 of the INS 8250 provides a power-on reset function. The I/O port would normally occupy eight consecutive port addresses, but this is not possible because of the Timex/Sinclair port addressing scheme. Port addresses for the various INS 8250 registers are given in **table 1**.

Read, Write, I/O Request signals, and Address lines A2 and A5 are decoded by the 74LS138 to make the I/O R and I/O W signals. INS 8250 select signals are furnished by Address lines A0, A1, and A6. The remaining Address lines A3, A4, and A7 control the register addressing. One half of the 74LS74 is used to divide the 3.25 MHz computer clock by two, to make the 1.625 MHz clock for the INS 8250. The three 2N2222 transistors should provide adequate protection for this expensive chip.

construction of the I/O port

This unit can be built on an etched circuit board or on perfboard with point-to-point wiring. The finished unit is installed between the computer and the 16K memory pack. An edge connector with wire-wrap pins should be used to attach the I/O port to the computer expansion bus.

The edge connector is installed on the *foil* side of the board. Allow the pins to extend through the circuit board approximately 3/16 inch (4.76 mm). Bend the pins together slightly and solder them to a small extender board; this extender board duplicates the computer expansion bus and provides connections to the 16K memory pack.

Although not shown on the schematic, at least two capacitors of about 0.047 μF should be placed on the circuit board and connected from the +5 volt supply to ground for the purpose of filtering out switching transients. A 3.3 kilohm pull-up resistor may be required on the SOUT lead with some Terminal Units.

part of fig. 2

item	description
C1,C2,C3	
C8,C9,C14	0.1
C4,C5,C6,C7	0.01*
C10	0.0047
C11	10, 25V electrolytic
C12	0.047*
C13	1.0, 25V electrolytic
C15	1000, 35V electrolytic
CR1-CR4	1N914
CR5,CR8	5.1V zener
CR6,CR7	red LED
CR9	green LED
CR10-CR13	1N4002
P1	500 ohm panel-mount potentiometer
Q1	2N2907
Q2,Q3	2N2222
R1	10 kilohm
R2,R33	1.0 megohm
R3	2.0 megohm
R4	15 kilohm
R5,R13	270 kilohm
R6,R14,R31	430 kilohm
R7,R15,R32	520 kilohm
R8	75 kilohm
R9	36 kilohm
R10,R18	470 kilohm
R11	560
R12	150 kilohm
R16	62 kilohm
R17	30 kilohm
R19	300
R20	470
R21,R34,R35	620
R22	120 kilohm
R23,R24,	
R25,R26	22 kilohm
R27,R29	330 kilohm
R28	560 kilohm
R30	680 kilohm
R36,R37	5.1 kilohm
R38,R44	3.3 kilohm
R39	220 kilohm
R40	6.2 kilohm
R41	7.5 kilohm
R42	8.2 kilohm
R43	9.1 kilohm
R45	390
R46	1.5 kilohm
R101-R104,R110	2 kilohm
R105-R108	300 kilohm
R109	200 kilohm
S1	SPST switch
S2,S4	2PDT switch
S3	2P3POS wafer switch
T1	12.6V, 300 mA Radio Shack 273-1385
U1,U2	LM3900
U3	XR2206
U4	7812 volt reg (TC-220 case)
case	Radio Shack 270-252 or -272

*Denotes Mylar. Recommended these be 5 percent "V" series.
Digi-key part No. P4513 for the 0.01 μF
P4521 for the 0.047 μF
The layout on the PC board is for the pin spacing of 5/16 inch preset variable resistors:
Digi-key part No. K4A32 for 300 ohm
K4A23 for 2 kilohm
K4A25 for 200 kilohm
All other resistors 1/4 watt, 5 percent tolerance
Miscellaneous plugs, jecks and other hardware is necessary.
All capacitance is in microfarads.

(Values below are for low tone operation (see text).

R8	120 kilohm	R21	1.2 kilohm
R9	62 kilohm	R22	180 kilohm
R11	1.1 kilohm	R40	9.1 kilohm
R12	240 kilohm	R41	11 kilohm
R16	91 kilohm	R42	13 kilohm
R17	47 kilohm	R43	16 kilohm
R19	470	RV	510
R20	820	P1	1 kilohm

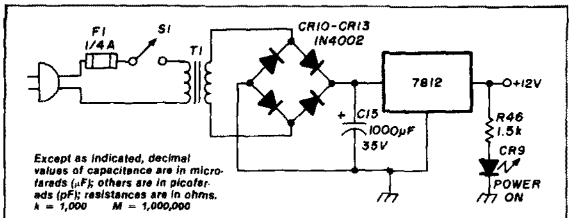


fig. 3. TU power supply schematic.

operation of the terminal unit

The schematic for the Terminal Unit is shown in **fig. 2. Figure 3** illustrates the circuit for the TU power supply. Wiring for the associated control and switching components is shown in **fig.** 4. UC 2A and 2B and their associated components make up the Mark filter which is tuned by R105. U2C and 2D and associated components comprise the Space filter. R106, R107, and R108 are switched to provide tuning for the 170 Hz, 425 Hz, and 850 Hz fixed shifts. A panel mounted control is switched into the circuit for the variable shift tuning.

Audio from the receiver speaker is fed into amplifier U1A and limited by diodes CR1 and CR2 to minimize effects of a building or fading signal. Good limiting action occurs with about 30 mV applied to the Audio In jack.

The limiter output is coupled to both active filters by C3. The outputs of the filters are rectified and applied to differential amplifier U1B. The signal from the Mark filter is applied to the noninverting input and the signal from the Space filter is applied to the inverting input. The output of U1B will, therefore, go high for a Mark and low for a Space. Capacitor C10 filters the individual tone frequencies from U1B's output.

Positive feedback applied to U1C "squares up" the output from the differential amplifier. U1D is an inverter for copying Normal and reverse signals. Driver transistors Q1 and Q2 provide power for the Space and Mark indicator LEDs CR6 and CR7. Zener diodes CR5 and CR8 furnish bias to keep the Mark indicator turned off when the output of U1B is less than approximately 6 volts and to keep the Space indicator off when the U1B ouput is above the 6 volt midpoint.

Function generator U3 provides the Mark and Space tones for transmit. Timing capacitor C12, together with the resistance connected to pin 7 or pin 8 determine the tone frequency. When pin 9 is low, frequency is controlled by the resistance connected to pin 8. When pin 9 is high, frequency is controlled at pin 7.

Variable resistor R101 tunes the Mark frequency. Variable resistors R102, R103, and R104 are switched to set the Space frequency for 170, 425, or 850 Hz. With the value shown for R38, the output level at pin 2 is about 0.2 volts. The output level can be increased by increasing the value of R38. A value of 50 kilohm will give an output of about one volt.

Frequencies of 2550 and 2975 are seriously attenuated in modern HF transceivers. If your operation is primarily on the HF bands you may prefer a Mark tone of 1275 and Space tones of 1445, 1700, and 2125 Hz. Component value changes for low tone operation are noted on the parts list.

constructing the terminal unit

While prototype was constructed on perfboard,

using the etched circuit board will save time and result in a neater unit. Install components, beginning with those that lie flat on the board and finishing with the power transformer. Refer to **fig. 4** for connecting the components that are not mounted on the circuit board.

I brought the 110 volt power line in from the back of the unit and mounted the fuse and jacks on the back of the case. The switches, LEDs, and the variable shift tuning control were installed on the front panel. Because it is necessary to watch both the Mark and Space LEDs when tuning in a signal, they should be mounted reasonably close together — not more than 2 inches apart.

Some older model transmitters use a tube to key the Press-To-Talk relay. In this case use a small reed relay (Radio Shack 275-233) connected between the collector of Q3 and +12 volts to key the transmitter Press-To-Talk circuit.

tuning and aligning the terminal unit

Be sure to do the following before installing the integrated circuits or applying power:

• Check for shorts or any abnormally low resistance reading across C15.

• Recheck the polarity of all diodes, electrolytic capacitors, and proper orientation of transistors.

• Turn on the power.

• Confirm that the Power On LED lights up.

• Check for 12 volts at the proper pins of each socket.

Then turn off the power and install the ICs, observing proper pin 1 orientation. Turn on the power and check the ICs, electrolytic capacitors, and the voltage regulator for heat. Although the voltage regulator may be slightly warm, nothing should feel hot.

Ground the Audio In jack and observe the Mark and Space indicators. It should be possible to turn on either LED by adjustment of R109; adjust R109 to turn them off. In darkness, it should be possible to see a faint glow in both LEDs; use R109 to balance this indication.

Remove the ground from the Audio in jack and place it on the FSK In jack to simulate a Mark. Connect a frequency counter to the AFSK Out jack. Adjust output level control R110, if necessary, for a stable reading on the counter. Adjust R101 for a frequency reading of 2125 Hz.

Remove the ground from the FSK In jack, set S3 to N (narrow shift), and adjust R102 for a reading of 2295 Hz. Set S3 to M and adjust R103 for a reading of 2550 Hz. With S3 in W, adjust R104 for a frequency reading of 2975 Hz.

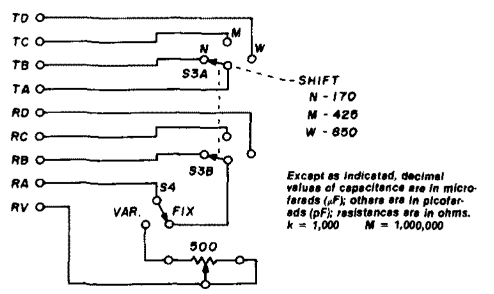


fig. 4. TU controls schematic.

Ground the FSK In jack again, and connect the AFSK Out jack to the Audio In jack. Tune the Mark filter to resonance by adjusting R105 for maximum brilliance of the Mark LED. If necessary, reduce the AFSK output level with R110 to achieve a sharp tuning peak.

Set switch S4 to FIX. Remove the ground from the FSK In jack, set S3 to N, and adjust R106 for maximum brilliance of the Space LED. With S3 set to M, adjust R107 for maximum brilliance again of the Space LED. Repeat this adjustment with S3 set to W using R108. Place S4 to VAR and S3 to M.

Adjust the variable shift tuning control for maximum brilliance of the Space LED. Repeat this check in N and in W positions. If the knob on the variable tuning control is set at the 12 o'clock position for 425 Hz shift, the 170 Hz tuning will be at about the 8:30 position and the 850 Hz tuning will be at about the 3:30 position.

Ground the FSK In jack once again. Set R110 at about its midpoint and place S2 to M. Measure the output of the Mark filter at the anode of CR3. Remove the ground from the FSK In jack and measure the output of the Space filter at the anode of CR4.

If the output of either filter is less than 8 volts peak-to-peak or 2.8 volts RMS at resonance, decrease R5 in the case of the Mark filter or R13 in the case of the Space filter, to a value of 240 kilohms. It should not be necessary to go below 240 kilohms if 5 percent or better tolerance parts have been used.

software notes

Table 1 shows example routines for initializing the port, reading data from the port, and sending data to the port. The formula for determining the Divisor is:

$$Divisor = \frac{1.625 \times 10^6}{baud\ rate \times 16}$$

For 60 WPM RTTY (45.45 Baud):

$$\frac{1.625 \times 10^6}{45.45 \times 16} \cong 2235$$

The Divisor Most Significant Byte is (in BASIC):

LET DMSB = INT (DIVISOR/256)

The Divisor Least Significant Byte is:

*LET DLSB = DIVISOR − DMSB * 256*

For 45.45 Baud:

Divisor Least Significant Byte = 187

Divisor Most Significant Byte = 8

The port address for the Line Control register is AF(hex). When bit 7 of this register is set, the Divisor Latches are enabled and port address 27(hex) allows the Divisor Least Significant Byte to be entered. Address A7(hex) allows the Divisor Most Significant Byte to be entered. When bit 7 of the Line Control register is reset, port 27 (hex) addresses the Transmit and Receive buffers. Line Control register bits 0 and 1 control the number of data bits per character. Bit 2 determines the number of stop bits.

initializing the I/O port

1. Set bit 7 of Line Control register to gain access to the Divisor Latches.

2. Enter Divisor bytes.

3. Reset bit 7 of Line Control register and set up for desired mode.

4. Disable interrupts or set up for desired interrupt mode.

5. Set any desired output conditions with the Modem Control register.

reading I/O port

1. Read the Line Status register. Bit 0 will be set if a character is ready in the Receive buffer.

2. If a character is ready, read it from the Receive buffer. Bit 0 in the Line Status register will automatically be reset when the character is read.

writing to the I/O port

1. Read the Line Status register. Bit 5 will be set if the port is ready for a new character.

2. If the port is ready, write the character to the Transmit buffer.

The INS 8250 I/O port provides a simple solution for obtaining stable, software-selectable baud rates. I think the Terminal Unit design represents a reasonable compromise between circuit complexity and performance. Notes have been given on software design to help those interested gain a better understanding of Input/Output programming.

ham radio

run RTTY
on your VIC-20

Interface allows VIC-20 to command transceiver for CW and RTTY operation

It all started at a meeting of our ham club, when KB4SM sold me a Kantronics *Hamsoft*™ cartridge (He had just upgraded to *Hamtext*™.) So I plugged *Hamsoft* into my VIC-20 and rigged my key to see how the unit would print my CW. I found it to be very critical of my fist, with such aberrations as "the" coming out as "6E." After rigging a diode to rectify the receiver audio, I found that it was just as critical of the other fellow's keying, so I felt better.

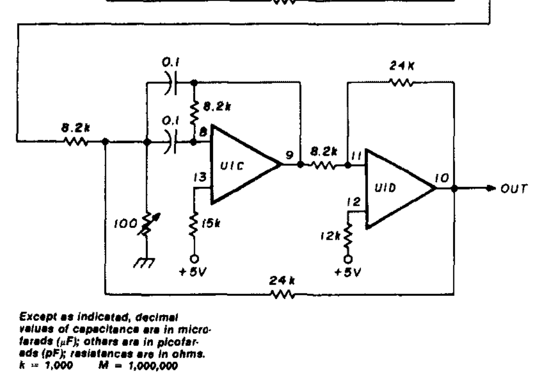


fig. 1. Schematic of bandpass filter (from W4AYV, *ham radio*, April, 1979, page 47.

Next came the RTTY, which was the main purpose of the project. I used an LM567 chip, and sure enough, it printed out some of the stuff that I tuned in. At this point it became only too obvious that extra selectivity was essential, so I looked in some back issues of *ham radio* and found an active bandpass filter design[1] using an LM3900 chip that I had picked up at the Memphis Hamfest (**fig. 1**). With these two units combined, reception was quite tolerable. I hooked a pair of high-impedance headphones at the output of the bandpass filter to aid in tuning in the signal.

The next consideration was transmission of RTTY and CW. National Semiconductor's *Linear Handbook* provided a circuit for the AFSK oscillator. The output was found to be a square wave of several volts, peak-to-peak. This was far too much signal to substitute for the usual microphone output, so a three-section RC filter was introduced. In order to load the RC oscillator as lightly as possible, the input resistor was chosen to be about 39 kilohms, with the others being 12 kilohms. (These may be increased if the audio output should still be somewhat high.) Capacitors were all 0.01 μF ceramic. The output was now a nice looking sine wave, necessary for a clean signal.

circuit adjustments

To vary the frequency of the AFSK oscillator, a smaller capacitor is switched in and out of the frequency-determining circuit by a 2N2222 transistor. When used in this function, no DC voltage is applied to the collector of the transistor.

The AFSK oscillator frequency is given as the reciprocal of the RC product. A capacitor of 0.068 μF was combined with a resistor of 6.8 kilohms. The smaller capacitor that is switched for frequency

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change is 0.0056 µF. If you have a frequency counter available, the AFSK oscillator output may be established and adjusted, if necessary. It is more important to obtain a frequency shift of 170 Hz, when using an SSB transceiver, because corrections are automatic as the signal is tuned in.

The frequency of the bandpass filter is now adjusted to peak at the frequency of the AFSK oscillator by means of the two 100-ohm trim pots.

Next, the position of the potentiometer controlling the PLL frequency of the decoder, when its frequency coincides with that of the AFSK oscillator, must be determined. This is done by applying DC to both chips and either comparing the signals on a scope or combining them through a temporary resistive network, and adjusting to zero beat. This is the normal operating position. Any minor differences between your frequency and that of the other station will be compensated with this control, so that your transmitting frequency will not be affected.

For CW operation, a PNP transistor operates a reed relay that keys the transmitter directly. Although the transistor might do the job without the relay, I did not want to take a chance that some problem with the transmitter might damage the computer.

space and mark signals

The decoder chosen makes use only of the mark signal and ignores the space signal. Although marginal signals may better be handled by a decoder that makes use of both components, this is a satisfactory arrange-

table 1. Parts list.

quantity	description
1	LM3900 quad amplifier
2	LM567 tone decoders
1	14-pin socket
2	8-pin sockets
5	0.1 µF Mylar capacitors
5	0.01 µF ceramic disc capacitors
1	0.1 µF ceramic disc capacitor (can use Mylar
2	1.0µF tantalum capacitors
1	0.068 Mylar capacitor
1	0.0056 Mylar capacitor
2	1 kilohm 1/4-watt resistors
1	3 kilohm 1/4-watt resistor
6	8.2 kilohm 1/4-watt resistors
4	24 kilohm 1/4-watt resistors
4	12 kilohm 1/4-watt resistors
2	15 kilohm 1/4-watt resistors
1	6.8 kilohm 1/4-watt resistor
1	2N2222 transistor
1	2N3906 transistor
1	5-volt reed relay
1	LED
1	DPDT slide or toggle switch
3	miniature 2-circuit jacks
1	miniature 3-circuit jack
1	5 kilohm linear potentiometer
1	4 × 5 inch PC board, copper on one side
1	18-pin edge connector (optional)

suitable housing box for interface

appropriate cables and connectors to mate with transceiver

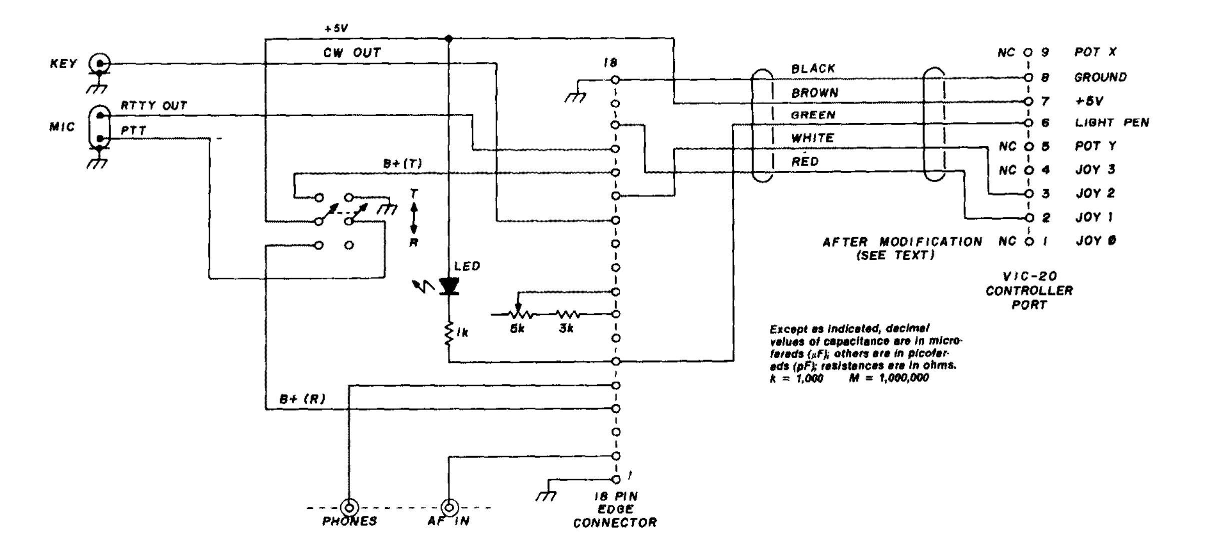

fig. 2. Additional interface wiring.

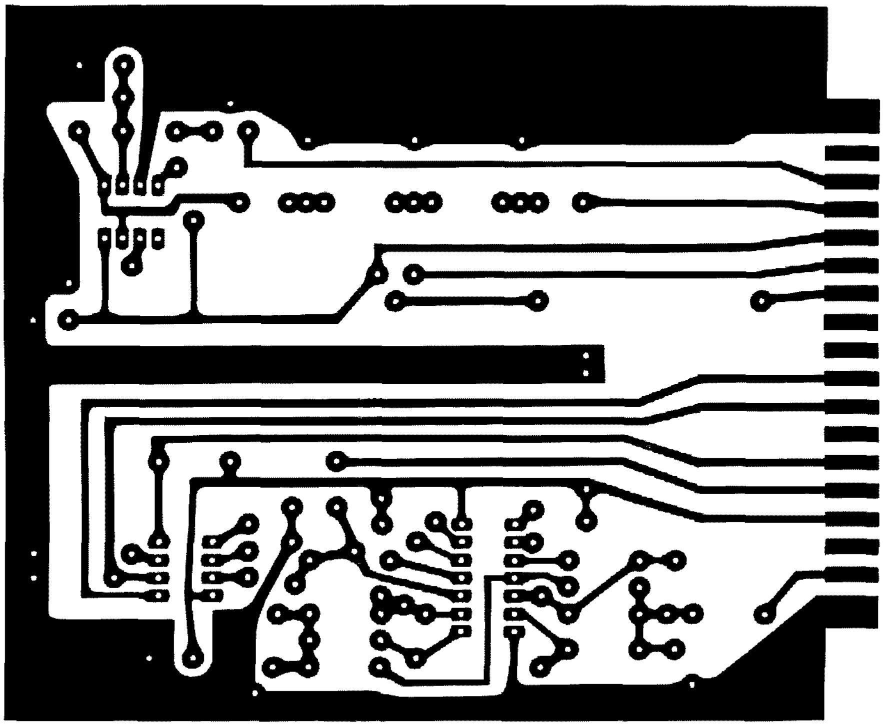

fig. 3. Foil side of PC board for the CW/RTTY.

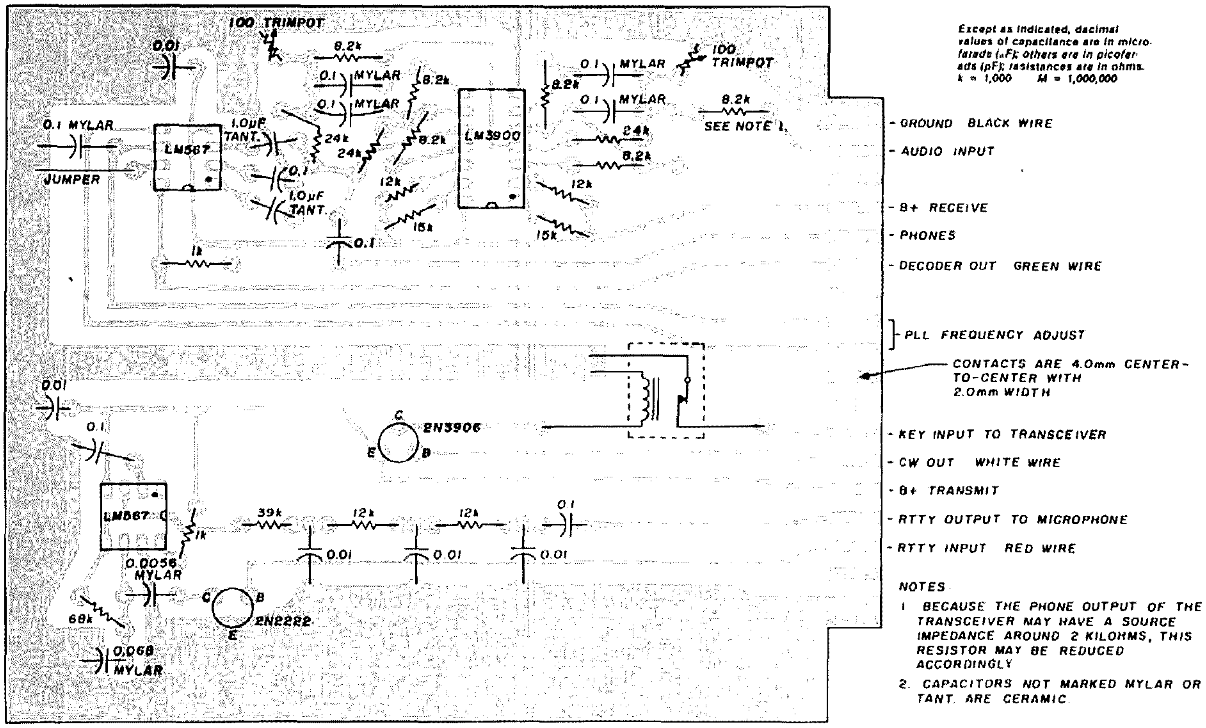

fig. 4. Component side of PC board for the CW/RTTY interface.

ment; signals that read R1 on the meter will usually give good print. A minor advantage appears with this decoder when widely different frequency shifts are encountered, because they present no problem to the mark-only decoder.

Most RTTY transmission and reception is on the LSB mode of the transceiver. In transceivers with provisions in the IF channel for a CW filter, this filter will usually be found to be tied into the USB mode. A number of stations operate this way, with inverted signals. Reception is easily corrected by adjusting the PLL frequency.

Accommodating transmission to the USB mode may be done in several ways. A friend assures me that if I press the K button instead of T, when activating the system, the transmission will not be inverted when on the USB mode. I haven't tried it, so I'm not sure. Another way would be to locate a transistor inverter stage between the decoder output and the computer light pen input terminal, and another in front of the NPN keyer transistor.

power supply notes

A separate power supply was originallly used for the interface, but one day the bandpass filter as well as the decoder and AFSK oscillator were run from a 5-volt common supply. Because little if any difference could be detected, 5 volts was used for the entire interface. The importance of this finding was that now the power to the interface could be supplied by the computer eliminating a separate DC source. The VIC-20 can supply +5 volts at up to 100 mA; the interface requires about 20 mA.

The cable that comes with the *Hamsoft* cartridge has five wires: black for ground, green for demodulator input, white for CW output, red for RTTY output, and brown for RTTY output (inverted). This last capability was exchanged for +5 volts supply, by removing the pin from hole No. 1 and moving it to hole No. 7. This probably eliminates the previously mentioned possibility of inverting the signal with the K for T change. However, the elimination of a separate power supply is well worth the effort.

additional wiring

Additional wiring included a DPDT slide switch for transmit or receive. The +5 volts is switched from the bandpass filter and decoder on receive to the AFSK oscillator on transmit. Also, the other section of the switch grounds the PTT line to the microphone connector of the transceiver. An LED is connected from the decoder output to the +5 volts, through a limiting 1 kilohm resistor. All of this extra wiring is shown in fig. 2.

The system also receives ASCII quite well. PLL adjustment on ASCII seems a bit more critical than on RTTY, but copy seems every bit as good. I have not tried ASCII on transmit because the slowest RTTY speed is still too fast for my typing ability.

The resistors used in the system are 5 percent. Any *frequency* determining capacitors should be Mylar or the equivalent. Ceramics are OK for bandpass, but I used the little Tantalum type wherever possible.

The system was mounted on a PC board, arranged to plug into an 18-pin edge connector, mounted in a 4 × 6-inch (10 × 15 cm) console-type box. Any PC board arrangement should do; perf board would be a bit messy.

The PC board as seen from both sides is shown in **figs. 3** and **4**. A parts list is provided in **table 1**. The reed relay as mounted on the PC board is shown in schematic form because these items vary considerably, and because I believe some freedom should be left to the individual builder.

Hamsoft™ and *Hamtext*™ are manufactured by Kantronics, Inc., 1202 E. 23rd Street, Lawrence, Kansas 66044.

references

1. *Linear Handbook*, National Semiconductor, 2900 Semiconductor Drive, Santa Clara, California 95051.
2. Nat Stinnetee, W4AYV, "Active Bandpass Filter for RTTY," *ham radio*, April, 1979, page 46.

ham radio

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VHF/UHF WORLD

Joe Reisert W1JR

stacking antennas: part 1

There are two basic ways to obtain high antenna gain in the VHF/UHF spectrum. Either you build a high-gain antenna with a single feed (such as a parabolic dish or long Yagi), or you build a number of single-feed devices and array or stack them for higher gain. Parabolic dishes, which require only a single feed system, have been built with gains exceeding 60 dBi (dB above an isotropic radiator). (To convert dB over a dipole to dBi, add 2.15 dB.) Hence they are especially popular on EME, where high gain is necessary.

However, high gain dish type antennas can get quite large. For example, let's see what size dish would be necessary for EME. The minimum recommended antenna gain for 2-meter EME is 20 dBi. This would require a dish approximately 33 feet (10 meters) in diameter. The minimum recommended gain for 70 cm EME is 25 dBi, which would require an 18-foot (5.5 meter) diameter dish.[1] Furthermore, parabolic dishes are usually only 50 to 60 percent efficient and can present structural problems, especially for those locations where wind and snow are prevalent.

Yagis are replacing dishes

In recent years, the Yagi antenna has become very popular, particularly on 70 cm and lower frequencies. Its popularity is justly deserved because if properly designed, it can exceed 70 to 80 percent efficiency with only moderate wind load. A properly designed 20 dBi-gain Yagi would, however, require a boom length of about 13 wavelengths — 89 feet (27 meters) at 2 meters! A 30-foot (9 meter) 4.4 wavelength boom design is about the longest practical 2-meter Yagi, but it would have a gain of only about 16 to 17 dBi, 3 to 4 dB lower than desired for EME. Therefore, when high gains are required, two or more Yagi antennas are arrayed or stacked to obtain the required gain.

general principles in stacking

It is often said that every time you double the number of Yagis, you double (add 3 dB) the overall gain. It is also common to hear that the proper stacking distance for a Yagi is two thirds of the boom length. Are these statements true? *No.*

But don't lose heart. Since I'm all too often asked about proper stacking distance, I decided that it's time to update the material I've been distributing since my first talk on the subject (at the Central States VHF Conference in Kansas City in 1977) and present it here along with additional data.

This subject deserves more than just a set of charts or tables that quickly become obsolete as new designs appear. Also, there are many practical aspects of stacking that are often ignored. Therefore, I've decided to first discuss stacking concepts in depth.

Since the material required to thoroughly cover the subject of stacking is extensive, I've decided to devote two monthly columns to this topic. In part **1** (this month), I'll discuss the theory of stacking and provide the examples, tables, and charts required for a first cut. Part **2** (next month), will discuss the practical aspects of stacking and provide suggestions on how to properly use the material presented in part 1. At the conclusion of part 2, you should have all the necessary material to determine the proper stacking for any Yagi antenna of your choosing, even designs that are not yet available!

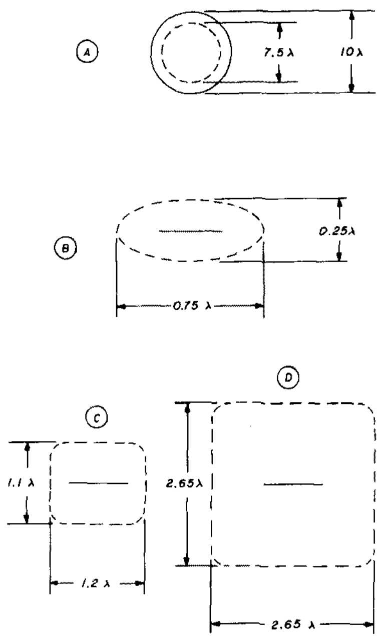


fig. 1. Typical effective apertures (shown within dotted lines) of some common antennas. ***(A)*** **is a 10 wavelength diameter parabolic dish;** ***(B)*** **is a half wavelength dipole;** ***(C)*** **is a 1 wavelength Yagi; and** ***(D)*** **is a 10 wavelength Yagi.**

table 1. Data on typical popular 2-meter through 23-cm Yagi designs including gain, boomlength, E and H beamwidths, E and H side lobe levels and recommended stacking distances. Data is believed to be accurate and has been gleaned from tests, data sheets, etc.

Yagi description	gain (dBi)	boom length (λ)	E B.W. degrees	H B.W. degrees	E S.L. (dB)	H S.L. (dB)	recommended stacking distance in E & H planes (λ)
general designs:							
NBS 3 element	9.25	0.4	57	72	23	12	1.00 × 0.60
NBS 5 element	11.35	0.8	48	56	22	13	1.20 × 0.90
NBS 6 element	12.35	1.2	40	42	19	12	1.40 × 1.10
NBS 12 element	14.40	2.2	34	36	17	13	1.55 × 1.40
NBS 17 element	15.55	3.2	28	33	16	12	1.80 × 1.35
NBS 15 element	16.35	4.2	26	29	17	13	1.95 × 1.75
2-meter designs:							
Cushcraft Jr. Boomer	14.40	2.2	34	36	17	13	1.55 × 1.40 (note 1)
Lunar 11 element	14.50*	2.6	31	34	17*	13*	1.65 × 1.50
F9FT 16 element	14.80	3.0	32	34	22	18*	1.80 × 1.50 (note 1)
KLM 2M-13LBA	15.00	3.1	28	33	18	15*	1.80 × 1.60 (note 1)
Cue Dee 15 element	15.15	3.1	30	32	16*	12*	1.70 × 1.40
Cushcraft Boomer	15.55	3.2	28	33	16	13	1.80 × 1.55
KLM 2M-16LBX	16.50	4.1	26	29	18*	15*	2.00 × 1.75 (note 1)
135 cm designs:							
Lunar 11 element	14.50	2.60	31	34	17*	13*	1.65 × 1.50
Cushcraft Boomer	16.35	4.2	26	29	17	13	1.95 × 1.75 (note 1)
KLM 220-22LBX	17.75	6.65	22	25	17	14	2.45 × 2.00 (note 1)
70-cm designs:							
KLM 432-16LB	15.20	5.3	30.0	33.0	17*	14*	1.70 × 1.55
K2RIW 13 element	15.40	5.3	29.5	29.5	10	7	1.50 × 1.50
K2RIW 19 element	17.35	5.6	24.0	26.0	18	15	2.10 × 1.80 (note 1)
F9FT 21 element	17.40	6.6	24.0	26.0	13	10*	2.10 × 1.75 (note 1)
FLEXA-YAGI 23 el.	17.95	7.2	24.0	25.0	17	15*	2.40 × 2.00
Cushcraft 424B	18.00	7.6	19.0	22.0	14	12	2.20 × 1.80 (note 1)
KLM 432-30LBX	19.40	9.6	19.0	20.0	17	14	2.70 × 2.40 (note 1)
W1JR 31 element	19.60	10.5	18.0	20.0	17	14*	2.70 × 2.40 (note 1)
23-cm designs:							
Tonna 23 element	17.00	7.5	19.0	19.0	12	10	2.40 × 2.40 (note 1)
W1JR 45 EL LPY	20.7	15.7	18.0	20.0	15	13*	2.85 × 2.65 (note 1)

*Estimated

Note 1. In this case actual tests have shown that a specific optimum is preferred (see text).

fundamental aperture concepts

Let us first examine some different antennas, each with its specific "effective aperture" or capture area. Some typical examples are shown in **fig. 1**. Other examples are contained in reference 2.

A 10 wavelength diameter parabolic antenna is shown in **fig. 1A**. It is easy to see how it has collection properties similar to that of the human ear. Note, however, that a dish antenna is not 100 percent efficient because it does not collect signals very well near its edge. Using simple geometry, the physical aperture of a 10 wavelength diameter dish is 78.5 square wavelengths, but its effective aperture is only approximately 44 square wavelengths.

A half-wave dipole antenna is shown in **fig. 1B**. Its aperture is more difficult to visualize. Note that its aperture extends out horizontally about 0.75 wavelength in the E plane and vertically about 0.25 wavelength in the H plane, bulging near the center and forming an ellipse. It has an effective aperture of approximately 0.13 square wavelength.

A Yagi has a slightly differently shaped aperture. A short (1 wavelength) conventional Yagi (one whose elements all lie completely in the same plane) is shown in **fig. 1C** and has a somewhat rectangular aperture, being wider in the E (horizontal) plane than in the H (vertical) plane. Its aperture is approximately 1.3 square wavelengths . A properly designed conventional 10 wavelength long Yagi as seen in **fig. 1D** has an almost square aperture of approximately 7 square wavelengths.

*Additional recommended reading: *Significant Phased Array Papers*, edited by R.C. Hansen, PN 0-89006-019-3, Artech House, Inc., 610 Washington Street, Dedham, Massachusetts 02026 ($13.00 plus $2.50 postage and handling).

Some of you may want to research this subject further to see how I determined the apertures. If you know the individual antenna directivity gain[3] you can calculate the effective aperture using the following equation:

$$\text{effective aperture} = \frac{\text{gain}}{12.5} \qquad (1)$$

where effective aperture is in square wavelengths and gain is over isotropic as a numeric. For example, a ½ wavelength dipole has a gain of 1.64 (2.15 dBi). Therefore it has an effective aperture of 0.13 square wavelengths. A 1 wavelength Yagi has a directivity gain of approximately 16.5 (12.2 dBi) and an aperture of approximately 1.32 square wavelengths. Likewise, if we have a 10 wavelength Yagi with a gain of 87.5 (19.4 dBi), the aperture will be approximately 7 square wavelengths, quite an aperture increase over the 1 wavelength Yagi.

The above equation does not reveal the width or height of the aperture. In the case of the 10 wavelength Yagi the aperture is approximately square (**fig. 1D**). Therefore the horizontal and vertical dimensions are approximately the square root of the aperture ($\sqrt{7}$) or 2.65 wavelengths. Since the 1 wavelength Yagi is slightly rectangular (per **fig. 1C**), the aperture will be slightly wider than it is high or approximately 1.2 by 1.1 wavelengths, respectively.

Once the concept of aperture is understood, it is easy to see what happens when we try to stack two identical antennas. When they are in close proximity, their apertures overlap, as shown in **fig. 2A**. Hence the capture area will not be doubled. Also the gain will not be double (3 dB increase) that of the single antenna. Furthermore, when identical antennas are too closely spaced, they introduce mutual impedance effects that can play strange games with the pattern, power distribution, and VSWR.

If we move the antennas apart until their apertures just touch, as shown in **fig. 2B**, we should produce almost twice the capture area (more on this later). Separating the two antennas further as shown in **fig. 2C**, will definitely double the capture area but is not desirable for reasons to be discussed shortly. Similarly, stacking in the vertical plane is also possible and yields a similar increase in gain.

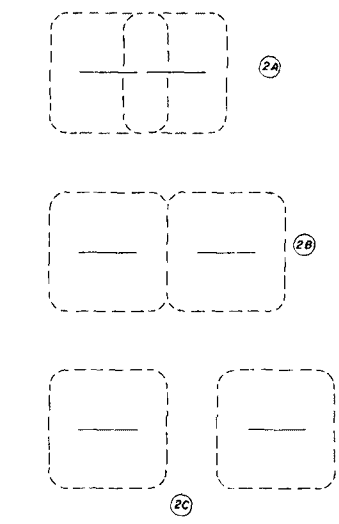


fig. 2. Effective apertures when stacking two square aperture antennas. *(A)* is too closely spaced; *(B)* is optimum spacing; and *(C)* is too loosely spaced.

stacking patterns

"So," you ask, "how does this relate to my Yagi?" First let us see what happens by looking at some typical antenna patterns. **Figure 3A** illustrates a typical antenna pattern for a 3-element Yagi.[4] Note that the half-power beamwidth is approximately 80 degrees and that this antenna pattern has very low side lobes.

We will stack two identical 3-element Yagis close together as shown in **fig. 2A**. The resultant antenna pattern is shown in **fig. 3B**. Note that the main beam narrows to about 50 degrees and the pattern is still very clean.

Next, let's separate the antennas further apart as shown in **fig. 2B**. The *resultant antenna pattern* is shown in **fig. 3C**. Note that the pattern beamwidth has become even narrower, 40 degrees, half the beamwidth of the original antenna. Also note that significant new lobes appear. These are properly referred to as "grating lobes" to differentiate them from the original single antenna's side lobes. The grating lobes in **fig. 3C** are only about 13 dB below the main beam. This separation is considered the optimum stacking distance.[5]

In **fig. 2C**, the antennas are spaced much further apart. The resulting antenna pattern is shown in **fig. 3D**. Note that in this case the main beam is approximately 20 degrees wide, or 25 percent that of the original antenna being stacked. Also note that the number of the grating lobes has increased to four, with maximum amplitude only 2 dB below the main beam. This is a technique often used by radio astronomers in interferometry. It allows very narrow beamwidths for greater accuracy in determining the position of extra-terrestrial objects. However, it is not usually desirable for Amateurs!

The patterns just shown all came from a "clean" Yagi. *If you look closely you will see that all the grating lobes and nulls were formed from within the area of the original antenna pattern.* This point is stressed because most Yagi antennas, especially those that are 1 wavelength or longer, usually have many side lobes *before* being stacked. The more side lobes you start with, the greater the chances are that the resulting pattern will be much "dirtier" and more complex than desired. Suppression of grating lobes is a difficult, if not impossible, task. Therefore, the real limitations when stacking antennas are the beamwidth, the side lobes in the antenna to be stacked, and the allowable level of the grating lobes. Despite the appearance that grating lobes are "robbing power" from the main beam, in actuality they are not since each grating lobe is very narrow. However, grating lobes are sources of extraneous noise or extra signal pickup, a killer on EME, and when there is lots of QRM. Incidentally, if you have many strong grating lobes, it is easy to accidentally peak your antenna on one of them instead of on the main beam!

To review, the optimum stacking distance for two identical antennas oc-

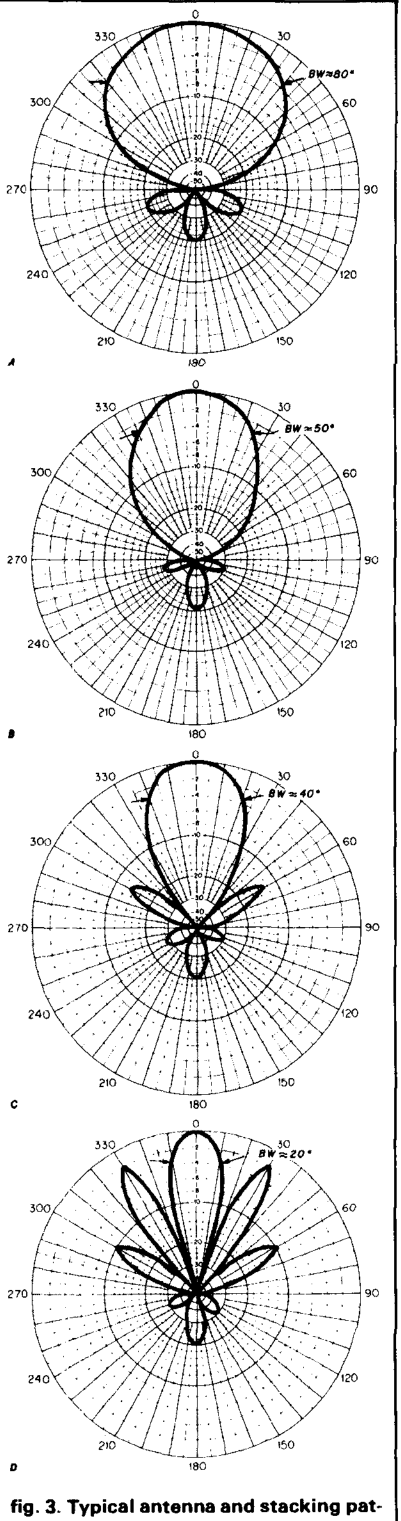
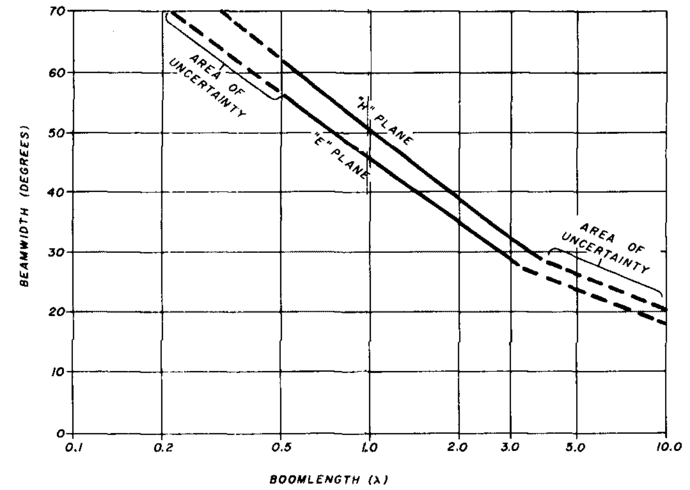


fig. 3. Typical antenna and stacking patterns. *(A)* is a typical 3-element Yagi; *(B)* two 3-element Yagis stacked 1/2 wavelength; *(C)* two 3-element Yagis stacked 3/4 wavelength; and *(D)* two 3-element Yagis stacked 2 wavelengths.



4. Typical beamwidths versus boomlength for well-designed Yagi antennas.

curs when the beamwidth of the array has been narrowed to about 50 percent and the grating lobes are approximately 13 dB below the main beam (more on this later). If four antennas are stacked in the same plane, optimum stacking would yield about 25 percent of the beamwidth of the original antenna while the grating lobes should still be 13 dB below the main beam.

actual stacking

There are two basic stacking methods: uniformly illuminated and shaped. Uniformly illuminated means that the antennas are all spaced the same distance apart in each plane and each is fed with the same amount of power. This method yields the maximum gain for its size and is the method most often used by Amateurs.

The shaped method is often used by professional antenna designers, especially in phased-array radars, where very low grating lobes are necessary. The individual antennas may be unequally spaced (sometimes one is completely left out!) and often are fed with different amounts of power. Since these techniques yield lower gain and are quite complex, they are usually not desired by Amateurs.

The subject that I have been discussing is called "pattern multiplication."[6] For more information on these techniques (pattern multiplication), see references 5 through 10.*

The most important parameters needed to determine optimum stacking distance are the beamwidths in the E and H plane of the antenna to be stacked. Also, the level of the first side lobe on the antennas to be stacked is important.

Most antenna manufacturers and antenna designers know the E plane beamwidth of their antenna very accurately since it is not difficult to measure. As discussed in reference 3, the beamwidths are often specified and accurate (in contrast to the gain claims). If the beamwidths are not known, they can be estimated from the "true" antenna gain. Several gain determining methods were described in detail in last May's column.[3] To save you even further time, I have prepared **table 1**, which lists many parameters of some of the most popular Yagi antennas.

The beamwidths of a Yagi antenna can be estimated if the boomlength is known. To assist you in this exercise, I have prepared a graph of E and H beamwidths versus boomlength for typical Yagi antennas (see **fig. 4**). All you need to know is the boomlength in wavelengths. For example, the E and H plane beamwidths of a typical

2 wavelength Yagi are approximately 35 and 39 degrees, respectively.

Also note in **fig. 4** that in a conventional Yagi the E plane is typically narrower than the H plane. While the E plane beamwidth is often available on data sheets, the H plane beamwidth is seldom shown (it is slightly more difficult to accurately measure). However, it can usually be "guesstimated" to be 10 percent greater than the E plane. For example, a Yagi with a 30-degree E plane beamwidth has a typical H plane beamwidth of 33 degrees.

Also, many Yagi antennas have side lobes that are so strong that they are equal to or greater than the idealized grating lobe desired after stacking! Therefore, it should be obvious that if the sidelobes on an antenna to be stacked are equal to or less than 13 dB below the main beam peak, they can't be optimally stacked. In this case the antennas must be placed closer together than optimum and consequently will yield lower stacking gain (more on this later)!

As a rule of thumb, the H plane side lobe level on a conventional Yagi is typically 3 dB stronger than the E plane lobe. Therefore, a Yagi with an 18 dB down sidelobe in the E plane probably has an H plane side lobe approximately 15 dB down from the main beam.

Now that we have determined the beamwidth and side lobe levels of our antenna, how do we determine the optimum stacking distance? For antennas with very low side lobes (at least 18 dB below the main beam):

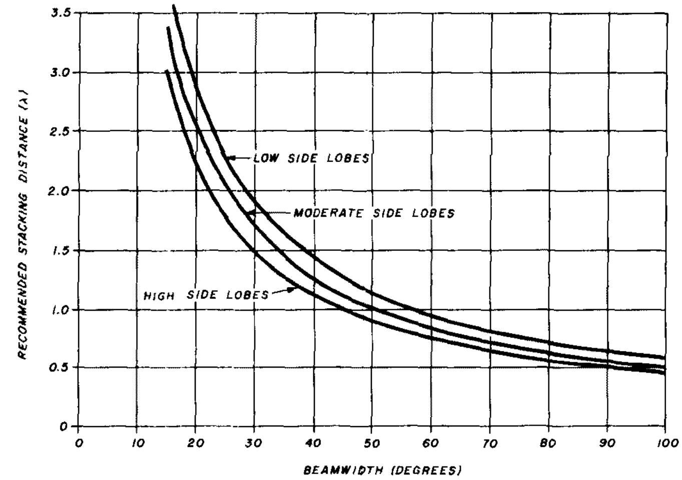
$$stacking\ distance \approx \frac{57}{beamwidth} \quad \textbf{(2)}$$

where stacking distance is in wavelength and beamwidth is in degrees.[10]

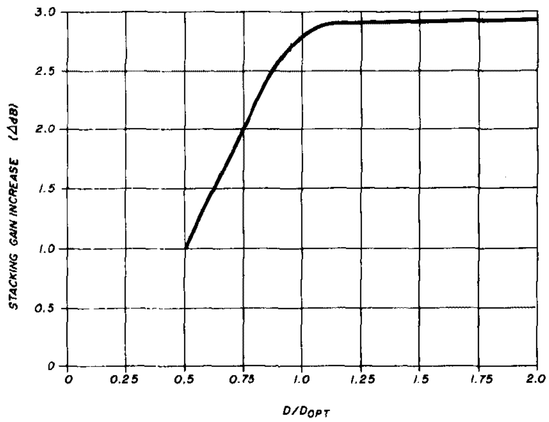
If the side lobes are typically 13-17 dB down, the usual situation, my tests have verified that:

$$stacking\ distance \approx \frac{51}{beamwidth} \quad \textbf{(3)}$$

If the side lobes are only 12 dB down (or less) (typical of the H plane of many Yagis):



5. Recommended stacking distance from *eqs. 2, 3,* and *4*. See text for proper usage and interpolation.



6. Typical gain increase due to stacking antennas. "D" is the stacking distance and "Dopt" is the recommended optimum from *eqs. 2, 3,* and *4*.

$$stacking\ distance \approx \frac{45}{beamwidth} \quad \textbf{(4)}$$

For simplicity, I have incorporated these formulas into a graph **fig. 5**), which is an updated version of the graph discussed earlier. It incorporates the recommended stacking for antennas with different levels of side lobes.

It can also be seen that there is room for interpolation using **eqs. 2, 3,** and **4** if desired.

For example, the NBS 17 element 3.2 wavelength Yagi has an E plane beamwidth of 28 degrees and a 16 dB down side lobe. Therefore use **eq. 3** (or the graph). Hence, the recommended E plane stacking distance is approximately 1.8 wavelength. The H plane beamwidth is 33 degrees but the side lobe is only 12 dB down. Using **eq. 4**, the recommended H plane stacking distance is approximately 1.35 wavelengths, quite a bit less than expected.

For those who do not want to make the required calculations, **table 1** also includes the recommended stacking distance for the antennas listed. In some cases, actual test measurements have been taken to determine the optimum spacing. Therefore, if the recommended stacking distance is different from that which you calculate

(signified by note 1), it is the preferred value since actual tests have verified its validity.

stacking gain

So how much gain do you get if you use the recommended stacking distance? Günther Hoch, DL6WU, has carried out tests and presented some answers to this question.[10] I have incorporated his information in **fig. 6.** It can be seen that the gain approaches 3 dB, but only at very wide spacing as discussed earlier. A typical optimum value is about 2.8 dB. If you understack (i.e., position too closely) by about 25 percent (as illustrated in **fig. 2A**), the gain increase (from a single antenna to the array) will be reduced to about 2 dB and there will be almost no grating lobes! Obviously there is not a great degree of freedom when optimum gain and pattern are concerned.

Finally when an antenna has high side lobes, it must be stacked closer to control the grating lobes as shown by **eq. 3** or **4**. This represents a form of understacking and lower gain. When this situation, plus feedline losses and mutual coupling, are considered, you are probably lucky to attain 2.5 dB even when the optimum stacking distance is used (more on this subject next month).

After studying this subject and reference 3, it will become obvious that the level of the first side lobe is a very important antenna parameter. Unfortunately, most antenna designers rarely list this parameter, but instead often list the worthless front-to-side ratio! A change in the literature indicating the level of the first side lobe would be an improvement.

final evaluation

You should be able to test your pattern using the information just given, especially if your antennas are stacked in the horizontal plane. The test methods described in reference 3 should be sufficient. For those on EME, the sun can be used as a rough check. Always remember that if the antenna beamwidth is narrower than originally calculated and/or the grating lobes are less than 13 dB down from the main lobe, you have probably overstacked (i.e., positioned your antennas too far apart).

summary

Part 1 of this two-part series has been written to give you a feel for what happens when two or more antennas are stacked. Typical examples have been provided along with the equations and graphs necessary for determining optimum stacking distance. New or improved Yagi antennas or those I may have failed to mention can be quickly evaluated using the information provided in this article. Exact stacking distance is not critical since there are many compromises, as discussed.

Part 2 of this article will delve a little deeper into the subject, emphasizing the practical aspects of the subject with recommendations on how to obtain optimal performance in your particular situation.

acknowledgements

I would particularly like to thank Günther Hoch, DL6WU, Dave Olean, K1WHS, and Steve Powlishen, K1FO, for the test data they shared with me while I was preparing this article.

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upcoming VHF/UHF events

April 5:	*EME Perigee*
April 15:	*ARRL 2-meter Sprint Contest*
April 21:	*1400 UTC: predicted peak of Lyrids meteor shower*
April 21:	*ARRL 135-cm Sprint Contest*
April 26-28:	*Dayton Hamvention VHF/UHF Program (contact WA8ONQ)*
May 1:	*ARRL 70-cm Sprint Contest*
May 4:	*EME Perigee*
May 4:	*1300 UTC: predicted peak of Eta Aquarids meteor shower*
May 3-5:	*West Coast VHF Conference (contact W6RXQ)*
May 9:	*ARRL 23-cm Sprint Contest*
May 18,19:	*Eastern VHF/UHF Conference (contact W1EJ)*
May 19:	*ARRL 6-meter Sprint Contest (tentative)*

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ATLAS 350XL owners group. Send QSL card with s/n your rig. Know anyone who repairs them? Have any technical information to share? Any questions? Rod, N5NM, Box 2169, Santa Fe, NM 87504.

CHASSIS and cabinet kits. SASE K3IWK.

HELP. Need schematic Atronics KB105MP CW keyboard. Company disappeared. Cavett, 8570 Herbert, Pennsauken, NJ 08109.

U.S. and **FOREIGN** Callbook, Radio Handbook and other ARRL books, 1950 to 1975 edition. Ham Radio, QST, 73, CQ magazines, second-hand for my private collection. Send price list to: Vittorio Bruni, Poste Succ. 8 — 05100 Terni Succ. 8, Italy.

IMRA, International Mission Radio Association, helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, NY 10538.

ANTIQUE RADIO Classified — The National Publication for Buyers and Sellers of old radios — published monthly. Sample copy $1.50. A.R.C., 9951-9 Sunrise, Cleveland, Ohio 44133.

WANTED: Old microphones, remote mixers other misc related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

AMATEUR RADIO CLASSES: Sponsored by the Hampden County Radio Association starting February 26, Agawam High School, 7 PM. Novice, Tech, General, Advanced or Extra. There is no charge for the classes. Text books must be purchased but are less than $10. Classes meet weekly for 10 weeks. For more information or to sign up contact Art Zavarella (413) 786-9115.

VE TEST SESSION: Sponsored by the Hampden County Radio Association, Saturday, May 18, 9 AM, Hampden-Wilbraham Regional High School,621 Main Street, Wilbraham, Mass. Exams for all license grades will be offered. Send completed FCC Form 610, a copy of current license and check for $4.00 payable to ARRL-VEC to: Yorke Phillips, K1BXE, 235 Ames Road, Hampden, MA 01036 prior to April 18, 1985.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 09224 (213) 774-1255.

RTTY-EXCLUSIVELY *for the Amateur Teleprinter. One year* $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

AMATEUR RADIO CLASSES: Sponsored by the 19-79 Amateur Radio Association. Evening classes start March 19 at the Chelsea (Mass) High School. Open to anyone interested in becoming a ham. Classes are free with a small cost for materials only. For information: 19-79 Amateur Radio Association, PO Box 171, Chelsea, MA 02150 include phone number.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details. W8UUR, 1025 Wildwood Road, Quincy, MI 49082.

DIGITAL AUTOMATIC DISPLAYS. All transceivers. Six 1/2" digits. 5" wide by 1-1/2" by 9" cabinet! Send $1.00 for information. Grand Systems, POB 2171, Blaine, Washington 98230. Canadians VE7LB.

Coming Events

ACTIVITIES

"Places to go..."

OHIO: Dayton Hamvention, April 26, 27, 28, Hara Arena and Exhibition Center, Dayton. Admission $8 advance, $10 at door. Good for all three days. Banquet $14 advance, $16 at door. Flea market space $17 in advance for all three days. Technical, ARRL and FCC forums. New products and exhibits. Special group meetings. YL forum. International VHF/UHF conference. Amateur of the Year Award. Special achievement awards. Pre-registration starts January 1, 1985. For further information; Dayton Amateur Radio Association, Box 44, Dayton, OH 45401 or phone (513) 433-7720.

OHIO: The 16th annual B*A*S*H will be held on the Friday night of the Dayton Hamvention, April 26 at the Convention Center, Main and Fifth Streets. Parking in adjacent city garage. Admission free to all. Sandwiches, snacks and COD bar available. Live entertainment. Two exciting awards and many others. For further information contact the Miami Valley FM Association, PO Box 263, Dayton, Ohio 45401.

ARIZONA: The Cochise Amateur Radio Association (CARA) invites you to the inauguration of the CARA Training Facility and Range, a 40 acre complex in Cochise County, 5 miles east of Sierra Vista on Mason Road. The dedication will be our annual Hamfest, May 4 and 5. A flea market is planned and all tailgaters are welcome. For information: The Cochise ARA, PO Box 1855, Sierra Vista, AZ 85636. Att: KB7HB.

WISCONSIN: The Madison Area Repeater Association (MARA) announces its 13th annual Madison Swapfest, Sunday, April 21, Dane County Exposition Center Forum Building in Madison. Doors open for commercial exhibitors and flea market sellers at 8 AM. General admission 9 AM. Everything for hams, computer hobbyists and experimenters. All-you-can-eat pancake breakfast and Bar-B-Q lunch will be available. Admission $2.50/advance; $3.00/door. Children 12 and under admitted free. Flea market tables $4.00 each/advance; $5.00/door. Reserve early. Talk in on WB9AER/R, 146.16/.76. For reservations or information: M.A.R.A., PO Box 3403, Madison WI 53704.

CALIFORNIA: The 43rd annual Fresno Hamfest, May 3, 4 and 5, Tropicana Lodge of Fresno, 1406 N. Blackstone. Program includes tech talks, swap tables and flea market, transmitter hunts, CW contest, ARRL forum, commercial exhibits, eyeball QSO's, buffet dinner and more. Registration $24.00 before April 19, 1985. $26.00 after that date. For information: Jane Price, WA6HSW, 2353 W. Simpson, Fresno, CA 93705.

ILLINOIS: The Kishwaukee Amateur Radio Club's annual Hamfest, May 5, DeKalb County Fairgrounds, Suydam Road, Sandwich. Donation $2.00 advance; $3.00 gate. Inside tables $5.00 each. Free parking. Outside areas for tailgaters. Overnight camping, no hookups. Coffee and donuts available for early birds. Food wagon thereafter. Talk in on 94, 13-73. For tickets write: K.A.R.C., Box 334, Sycamore, IL 60178. Our 30th Year!

COLORADO: The Grand Mesa Repeater Society's 6th annual Western Slope Amateur Radio and Computer Swapfest, Saturday, April 20, 10 AM to 4 PM. Location TBA in Grand Junction. Free admission. Swap tables $5.00 each. Indoor Swapfest, Amateur Radio exams, auction and refreshments. Talk in on 146.82 and 449.200. For tables or information: SASE to Larry Brooks, WB0ECV, 3185 Bunting Avenue, Grand Junction, CO 81504 or call (303) 434-5603.

SOUTH CAROLINA: The Blue Ridge Amateur Radio Society is sponsoring the 46th annual Hamfest and Electronic Flea Market, American Legion Fairgrounds in Greenville. Saturday, May 4 8 AM to 5 PM. Sunday, May 5 8 AM to 3 PM. Admission $3.00 advance and $4.00 at gate. VEC walk in exams, Wouff Hong ceremony, ARRL State Convention, Saturday night banquet, ARES, QCWA, indoor dealer displays, indoor/outdoor flea market, food, beverages, snacks and camping. Early dealer/flea market setups with advance registration. For advance tickets and VEC exam info; Sue Chism, N4ENX, PO

Box 6751, Greenville, SC 29606. For additional information: Rancy Rice, WD4ADK, 1401 W. Parker Rd., Greenville, SC 29611.

OKLAHOMA: The Great Plains ARC will sponsor its 4th annual N.W. Oklahoma Eyeball & Swapmeet, Sunday, April 14, starting 9 AM in Mooreland. Admission $2.00. Dealer and swap tables available at no charge. VE tests given. Campsites available. Local airport. Covered dish dinner at noon. For further information: Gordon Richmond, NR5L, Rt. 1, Box 12, Mooreland, OK 73852 or Gerald Bowman, Box 356, Mooreland, OK 73852 or call (405) 994-5394. (405) 994-5453.

NORTH CAROLINA: The Raleigh Amateur Radio Society's 13th annual Hamfest, Sunday, April 14. 8 AM to 4 PM. NEW LOCATION Jim Graham Building, NC State Fairgrounds, Hillsborough Street, Raleigh. Pre-registration $3.50. $5.00 at the door. One flea market space, table, two chairs $5.00 (ours only please). Vendor setup Saturday 4-10 PM, Sunday 6-8 AM. Hamfest social Saturday night. Special interest meetings. Amateur FCC exams. Send Form 610 with copy of current license and check or MO for $4.00 to: W.C.A.R.S./V.E.C., Mr. John Johnson, WM4P, 2118 Lyndhurst Drive, Raleigh, NC 27610. CW and homebrew contests. Talk in on W4DW (146.04/146.64) and K4ITL (146.28/146.88). For further information: RARS Hamfest, PO Box 17124, Raleigh, NC 27619.

MICHIGAN: 1985 Blossomland Blast, Sunday, October 6, 1985. Write "Blast", PO Box 175, St. Joseph, MI 49085.

INDIANA: The Putnam County Amateur Radio Club's third Auction and Flea Market, April 6, Putnam County Fairgrounds, north of Greencastle on US 231. Doors open for setup at 0600. Flea Market 0800. Flea market tables $2.00 each. Admission $3.00. Children under 12 free. Auction starts 1300. Food and beverages available. Commercial exhibitors welcome. For information SASE to John S. Underwood, K9IIB, RFD 1, Box 10, Fillmore, IN 46128 or call (317) 246-6335.

NEW YORK: Indoor/outdoor Flea Market sponsored by the Suffolk County Radio Club, Sunday, May 5, 8 AM to 3 PM, Republic Lodge No. 1987, 585 Broadhollow Road (Rt. 110), Melville. Refreshments available. Free parking. Admission $2.00. (Spouse and kids free). Indoor tables $7.00, outdoor space $5.00, includes one admission. Talk in on 144.61/145.21 and 146.52. For information: Richard Tygar, AC2P. (516) 643-5956 evenings.

MISSOURI: The PHD Amateur Radio Association is sponsoring the State ARRL Convention, Saturday and Sunday, April 13 and 14, Trade Mart Building II, Kansas City Downtown Airport. Doors open 9:30-5:30 both days. Commercial setput 7-9 PM Friday, 7-9 AM Saturday. Saturday evening banquet. Special guest: Dale Clift, WA3NLO, ARRL General Manager. Forums include ARRL, computer, FCC, VE, DX, PR, CW contest, homebrew contest and much more. Exams by PHDVEC Friday 5 to 7 PM, Saturday and Sunday, 8 AM. Send application with $1.00 and SASE to PHDVEC, PO Box 11, Liberty, MO 64068-0011 by April 8, 1985. Registration $4.00 for both days. Banquet $10.50. Swap tables $10.00 for both days, includes one registration per table. Free parking. RV's welcome but no hookups. Talk in on 146.34/94. Send registrations to: PHD ARA, PO Box 11, Liberty, MO 64068-0011. Phone (816) 781-7313 or 452-9321.

ARKANSAS: The Northwest Arkansas ARC will hold its 5th annual Hamfest/Swapfest, Saturday, May 4, Rogers Youth Center, 315 West Olive, Rogers. 8 AM to 4 PM. Setup 7 AM. Commercial/flea market tables $2.00. Admission free. Walk in FCC exams given 10 AM and 1:30 PM. Talk in on 16/76 and 52 simplex. For information SASE to: Ray Watson, N5HAP, 714 Maple Drive, Springdale, AR 72765 or Dave Perry, KE5QZ, 3201 N. 13th, Rogers, AR 72756.

CALIFORNIA: The 13th annual Sacramento Valley Amateur Radio Hamswap, Sunday, May 5, Placer County Fairgrounds, Roseville. 9 AM to 3 PM. Talk in on 145.190 and 224.78, K6IS repeaters. Free parking. For advance tables, tickets, information: Carl Schultz, KA6KWB, 2942 Goldenwyn Way, Rancho Cordova, CA 95670. (916) 336-9111.

MASSACHUSETTS: The Hampden County Radio Association Flea Market, Sunday, May 5, rain or shine, West Springfield Lodge of Elks, Morgan Road, West Springfield. 9 AM to 3 PM. Admission $1.00 per person. Tables $3.00 each. Dealers $3.00 per vehicle display. Food/refreshments available. Talk in on 147.105 up 600. For information: Paul Kress, WA1ZKT (413) 568-8291 or Steve Nelson, WA1EYF (413) 596-8216.

NEVADA: The Ziegfeld Showroom of the MGM Grand Hotel will host the first 1985 earth station industry Banquet, held in conjunction with the SPACE/STTI Las Vegas show, Monday evening, April 1 at 6 PM. Festivities include dinner, special guest speakers and presentations. All topped off with Jubilee, the MGM's musical extravaganza. Tickets $50.00 per person. Call SPACE (703) 549-6900. Credit cards accepted or mail check to 709 Pendleton Street, Alexandria, VA 22314 before March 15. Banquet tickets are not refundable.

OHIO: The Medina Two Meter Group is sponsoring a Hamfest, May 12, Medina County Community Center Building, Lafayette Rd., State Rt. 42 SW. 8 AM to 2 PM. Setup 7 AM. Refreshments and free parking. Tickets $3.00 advance, $3.50 at door. Tables $6.00. Flea market space $2.00. Talk in on 147.63/03, K8TV/R. For table reservations and tickets: PO Box 452, Medina, Ohio 44258 or (216) 725-5021.

LOUISIANA: BRARC Hamfest, May 11 and 12, Baton Rouge. Free admission. VE exams Saturday and Sunday, 30 day advance registration only. Send SASE, 610 and check for $4.00 to ARRL/VEC, George Perry, W5LVX, 17424 Lady Constance, Greenwell Springs, LA 70739. For further information SASE to Rick Pourciau, N5HHF, 879 Castle Kirk, Baton Rouge, LA 70808.

GEORGIA: The Athens Amateur Radio Club (formerly the N.E. Georgia ARC) will sponsor a Hamfest, April 21, 8:30 AM to 3:30 PM, Athens Vocational-Technical School, Highway 29, Athens. Registration is free. Talk in on 147.285. For information: Norman Archibald, KB4IIA, PO Box 225, Athens, GA 30603.

NEW MEXICO: The UNM and Westside ARC's are co-sponsoring a tailgate swapfest, April 20, 10 AM to 2 PM MST, UNM North Campus parking lot, Tucker Avenue and University Blvd., Albuquerque. There is no charge but bring your own tables. Talk in on 147.75/147.15 and 449.3/444.3 repeaters. For information SASE to Robert A. Scupp, WB5YYX, 648 Marquis Drive NE, Albuquerque, NM 87123. (505) 296-6546.

CALIFORNIA: Flea Market and FCC examinations. April 13, May 11, June 8, July 13, August 10 and September 14. Novice thru Extra exams given. Information call (408) 255-9000. Foothill College, Los Altos, CA. 73 Gordon, W6NLG VEC.

ILLINOIS: The Moultrie Amateur Radio Klub (MARK) Hamfest, Sunday, May 5, 8 AM to 3 PM, Moultrie County 4-H Center Fairgrounds, Cadwell Rd., 5 miles east of Sullivan. Heated indoor/large covered outdoor Flea Market. No charge to vendors. Vendors setup Saturday. No overnight hookups. Talk in on 655/055 and 52. For information: MARK, PO Box 79, Sullivan, IL 61951 or call Vernon Jack, K9SWY (217) 728-7596.

NEW HAMPSHIRE: Springfest '85, the 5th annual Flea Market/Hamfest, sponsored by the Great Bay Radio Association. Saturday, April 20, 9 AM to 3 PM, Somersworth Armory, Blackwater Road, Somersworth. Admission $1.00. Tables $8.00 includes one admission. Free parking. Food and refreshments available. Talk in on 146.40/147.00. For information/table reservations: Great Bay Radio Association, PO Box 911, Dover, NH 03820.

ILLINOIS: The Centralia Wireless Association's annual Hamfest, Sunday, May, Kaskaskia College Gymnasium, 3 miles NW of Centralia. Doors open 7 AM to setups. No charge for flea market and exhibit space. Some tables available. Admission to Hamfest is free. Food and refreshments available. Exams for all license classes except Novice will be given at 9 AM. Send completed Form 610, copy of current license and check for $4.00 payable to ARRL/VEC to Lou Hodges, W9IL, Route 1, Box 62A, Centralia, IL 62801 by April 5, 1985. For further information: David Conder, KA9QPC (618) 532-2772 or Lou Hodges, W9IL (618) 533-4724.

MASSACHUSETTS: The Framingham Amateur Radio Association's annual Spring Flea Market, Sunday, April 14, Framingham Civic League Bldg., 214 Concord St. (Rt. 126) downtown Framingham. Doors open 10 AM. Sellers setup begins 8:30. Admission $2.00. Tables $10.00 includes one free admission. Pre-registration required. Bargains galore! Contact Jon Weiner, K1VVC, 52 Overlook Drive, Framingham, MA 01701. (617) 877-7166.

MASSACHUSETTS: The Montachusett Amateur Radio Association's Flea Market, Saturday, April 27, Knights of Columbus Hall, Electric Avenue, Fitchburg. Doors open 9 AM to 3 PM. Dealer setup 8 AM. Admission $1.00. Tables $8.00 each. Refreshments available. Free parking. Talk in on 144.85/145.45 and 146.52 simplex. For tables send check payable to M.A.R.A., Jim Beauregard, KB1AY, 7 Mountain Avenue, Fitchburg, MA 01424.

MINNESOTA: The Arrowhead Radio Amateur Club announces "Swapfest '85", Saturday, May 11, Holiday Inn, 207 West Superior Street,, downtown Duluth. Doors open 8 AM for vendors. General admission 10 AM. Admission $4.00. 4' tables $5.00. Plenty of food. Free parking. Talk in on 146.34/94 repeater. For information: Bill Cossette, N0BKL, 15 Manitou Street, Duluth, MN 55808.

MINNESOTA: The Rochester Amateur Radio Club's 8th annual Hamfest, Saturday, April 20, John Adams Junior High School, 1525 NW 31st Street, Rochester. Doors open 8:30 AM. Large indoor flea market for radio and electronics items, refreshments and plenty of free parking. Talk in on 146.22/82. For further information: RARC, c/o WB0YEE, 2253 Nordic Ct., NW, Rochester, MN 55901.

OPERATING EVENTS

"Things to do..."

APRIL 13: Connecticut QSO Party sponsored by the Candlewood ARA, 1100Z 13 Apr to 1100Z 14 Apr, rest period 0500 to 1000Z. Send signal report, QSO number, ARRL section or Ct. County for stations worked inside CT. Club station W1QI counts 5 points per band/mode. Mail by 5 May 1985 (SASE for results) to CARA c/o R. Dillon, N2EFA, Box 143, Bethel, CT 06801.

MAY 18: ARMED FORCES DAY. The annual Armed Forces Day Communication Test. CW, SSB, RTTY and SSTV. Cross band contacts — military to Amateur cross band operations 18/1300 UTC to 19/0245 UTC May 1985. Military stations participating in cross band operations: Air, NMH, NPL, NAM, NMN, NZJ, NAV, NPG, WAR. Military stations will transmit on select frequencies and announce the specific Amateur band frequency being monitored. The CW and RTTY broadcasts will be a special Armed Forces Day message from the Secretary of Defense. Transcriptions of CW and/or RTTY receiving tests should be submitted "as received". Time, frequency and call sign of military station copied and name, call sign and address of individual submitting entry must be indicated. Entries must be postmarked no later than 25 May 1985. Send to following military commands: AIR — AFD Test, 2045CG/DONJM, Andrews AFB, DC 20331-5000. NAM, NAV, NPG — AFD Test, 4401 Massachusetts Ave., Washington, DC 20390-5290. WAR-AFD Test, Commander, USAISC, Att AS-OPS-CM, Ft. Huachuca, AZ 85613-5000.

APRIL 10, 11, 17, 18: All licensed women operators throughout the world are invited to participate in the DX-YL North American YL contest. DX YLs call "CQ North American YL" and NA YLs call "CQ DX YL". All bands may be used. Stations may be worked and counted once on each band and mode. Exchange station worked, QSO number, RS(T) state or country. For more information: Marty Silver, NY4H, 3118 Eton Road, Raleigh, NC 27608.

MAY 1, 2: Indiana Month of May Contest: Be the first to work 500 Indiana contacts. Exchange RST, state, province, or country, name and county (Indiana stations). Send copy of log, dup sheet and score sheet by June 30, 1985 to Russ Ryle, N9DHX, Southern Indiana QRP Group, PO Box 2466, Bloomington, IN 47402.

APRIL 30: Amateur Radio operators will have a rare opportunity to work an English Renaissance sailing ship when the Godspeed sails on a 10 week voyage from London, England to Jamestown, Virginia. The original Godspeed was one of three square-rigged vessels which brought the first permanent English settlers to the New World in the winter of 1606-7. Rigging the ship's radio systems has been coordinated by Neil Tanner, WA4CHQ. The Captain of the Godspeed is George Salley, KA4FVB. Special QSL cards have been designed. For more information contact Jamestown/Yorktown Fdn., PO Drawer JF, Williamsburg, VA 23187.

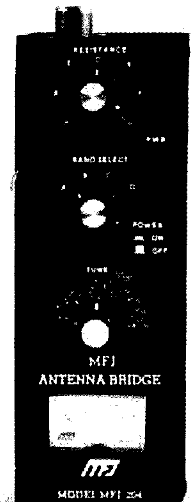
APRIL 20: Spring SSB Contest. 1200 UTC Saturday, to 2400 UTC Sunday. May operate a maximum of 24 hours. Exchanges: Members give RS, state/province/country and QRP ARCI membership number. Non-members give RS, state/province/country and power output. Stations may be worked once per band. Separate log sheets for each band must be received by May 21, 1985. Send logs to QRP ARCI Contest Chairman Eugene Smith, KA5NLY, PO Box 55010, Little Rock, AR 72225.

NEW products

antenna bridge

The new MFJ-204 Antenna Bridge lets you trim your antenna quickly and easily for its best performance.

The antenna bridge will give an accurate reading of your antenna resistance up to 500 ohms and will cover all the ham bands up to 30 MHz. When used to measure the resonant frequency of your antenna, it allows checking to see if the resonant frequency is higher or lower than desired. You can then lengthen or shorten your antenna based on the information gathered.



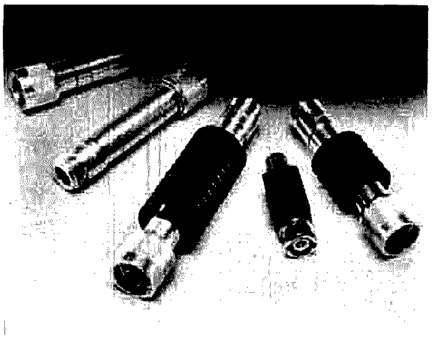
Priced at $79.95, plus $4.00 shipping, the MFJ-204 Antenna Bridge has a frequency counter jack for precise frequency measurement and can also be used as a signal generator. Housed in a sturdy black aluminum cabinet, the unit is very compact (4 × 2 × 2 inches) and operates on a single 9-volt battery or 110 VAC using MFJ's AC adapter (MFJ-1312, $9.95).

For further information, contact MFJ Enterprises, Inc., P.O. Box 494, Mississippi State, Mississippi 39762.

Circle #312 on Reader Service Card.

low/medium power coax attenuators

A series of 33 broadband, high-performance attenuators with six continuous-power ratings from 2 to 75 watts is now available from Bird Electronic Corporation. The 8300 series of 50-ohm fixed attenuators covers a frequency range from DC-4 GHz with a VSWR below 1.25, except for the 75-watt and the 2-watt units, which cover DC-2GHz. Models 8308 (75 watts), 8306 (25 watts), 8305 (15 watts), 8304 (10 watts), and 8303 (5 watts) are rated at a peak power of 3 kW with pulses to 5 microseconds wide, are available with attenuation levels of 3, 6, 10, 20, or 30 dB and have male N input/female N output connectors.



The 8300 series attenuators can be used in tandem for odd dB values, or in connection with Bird's TENULINE® high power attenuator (to 4000 watts) for additional attenuation.

For details, contact Bird Electronic Corporation, 30303 Aurora Road, Cleveland (Solon), Ohio 44139.

Circle #313 on Reader Service Card.

low noise amplifier

The R.L. Drake Company has announced the introduction of a new low-noise amplifier for consumer satellite television and other applications. The new Model 2574 provides a noise temperature better than 100 degrees Kelvin and utilizes a 15-volt DC power supply with an output of 3.7 to 4.2 GHz.

The price of the Model 2574 LNA is $195.00 (retail).

For further information, contact R.L. Drake, 540 Richard Street, Miamisburg, Ohio 45342.

STV filtering

Phantom Engineering, Inc. has introduced an improved version of their popular variable bandwidth filter, the IFP-1. The new IFP-1X replaces the IFP-1. It still features the fingertip IF bandwidth selection and IF gain control. The 70-MHz filter allows the bandwidth on the user's receiver to be progressively narrowed with the IFP-1X's four-position signal selector which, in cases of small dishes and/or terrestrial interference, has shown remarkable results.

The improvement comes from the ability of the IFP-1X to pass all control signals below 10 MHz that are used on quartz synthesized type receivers. Another improvement is the addition of the power supply with each unit at no additional cost.

For more information, contact Phantom Engineering, Inc., 16840 Joleen Way, Bldg. E, Morgan Hill, California 95037.

Circle #304 on Reader Service Card.

STV receiver

Luxor North America Corp.'s new Mark 2 remote-control STV receiver is a single integrated satellite component. With a built-in stereo processor and Dolby noise reduction, its advanced "block conversion" technology allows several TV sets in home, building, or neighborhood to share one STV antenna while enjoying independent channel selection.

Designed for use with regular, component TV, VCR, home or studio stereo, signal descrambler, dish positioning systems, and other audio-video equipment, Mark 2 is easily preprogrammed and operated by the infrared hand-held Remote Commander. An optional Remote Infrared Sensor allows use of the Commander in a room different from the receiver. An antenna positioning system can be remote-controlled. Digital channel and LED tuning indicators, and a signal strength meter, monitor receiver performance.

For details, contact Luxor North America Corp., P.O. Box 32, Bellevue, Washington 98009-0032.

Circle #315 on Reader Service Card.

tuning indicator kit

Heath Company has expanded its Amateur Radio line to include the HD-3006 Crossfire Tuning Indicator for quick and easy tuning of RTTY transmissions. Sixteen LEDs make up the HD-3006's visual display: eight vertical LEDs identify mark signal strength; eight horizontal LEDs indicate space signal strength. Tuning the HD-3006 for maximum vertical and horizontal display provides a strong signal for computers or RTTY printers. Each LED bar requires approximately 14 dB no-signal to signal voltage ratio for full operation. Minimum input signal is 0.3 VAC RMS or 0.5 VDC. Maximum signal is 15 VAC RMS or 15 VDC.

The HD-3006 Crossfire Tuning Indicator has a wide voltage range and is compatible with almost any interface/terminal unit that has oscilloscope outputs for tuning. The AC/DC cube-type power supply is included in the kit.

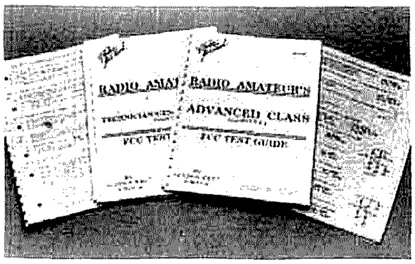
For information and a free catalog, contact Heath Company, Dept. 150-435, Benton Harbor,

Michigan 49022. In Canada, write Heath Company, 1020 Islington Avenue, Dept. 3100, Toronto, Ontario, M8Z 5Z3.

Circle #314 on Reader Service Card.

FCC/VE test guides

Test guides for every Amateur Radio class of license are now available from Gordon West's Radio School.



The test guides list all 500 test questions plus the multiple-choice answers in an attractive 8-1/2 × 11 inch manual. The questions, the distractors (wrong answers) and the right answers are listed exactly as they will be found on ARRL or W5YI volunteer examinations. The General and Advanced class test guides list 500 questions; the Extra class test guide, 400; and the Novice class test guide, 100.

Each test guide also includes study notes listing reference material from which the questions were derived and sources of further information about the answers. Formulas for solving the problems are also incorporated into each test guide.

Also included are several pages of instructions to the applicant on locating a Volunteer Exam Coordinator and how to sign up for a local volunteer-administered examination. Also included are the necessary test forms that applicants must fill out ahead of time, including the new FCC Form 610 (revised).

All test guides have been updated to reflect recent rewordings of FCC test questions. This will allow students to spot any format change in any of the FCC-approved questions.

Study guides are priced at $19.95 plus $3.00 postage. (Be sure to specify the license class you want.) Exclusive stereo Radio School 4-set cassette theory tapes are also available at $39.95; each set of four theory tapes includes Amateur Radio "sounds" that help illustrate specific questions on the examinations. (Be sure to specify the license class covered by the theory course you are requesting.)

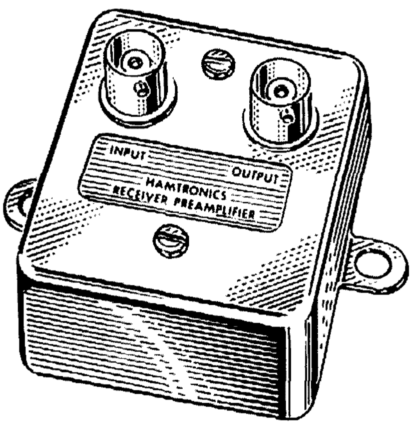
For more information on study guides, code and theory training tapes, and a colorful catalog on instructional materials for volunteers who give the Amateur Radio exams, contact Gordon West, RADIO SCHOOL, 2414 College Drive, Costa Mesa, California 92626.

Circle #302 on Reader Service Card.

NEW products

new Hamtronics® 800 MHz receiver preamplifier

A new 800-960 MHz version of its popular GaAs FET preamp is now available from Hamtronics, Inc. The LNG-800 preamp features a dual-gate GaAs MESFET with built-in diode protection against static discharge damage. The



unit has 11 dB of gain with a 1.5 dB noise figure. It is easy to install, operates on 13.6 VDC, and measures only 2 × 2 × 1-1/2 inches. The LNG series of GaAs FET preamps is available also for the high band and UHF band. Preamps in the LNG series, including the LNG-800, cost $49 plus $3 shipping.

For complete information on the LNG preamps, and other Hamtronics products for VHF and UHF, contact Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535.

Circle #107 on Reader Service Card.

J.I.L. SX-400 general coverage scanning monitor receiver

The new SX-400 general coverage scanning monitor receiver from J.I.L. Corporation is designed to cover AM/FM signals in the 26-520 MHz region. This frequency range should be of interest because of the unit's complete coverage of the FM/TV and the 220-MHz band as well as military frequencies above 225 MHz. A complete line of converters for expanding the SX-400's range from 100 kHz to 1.4 GHz — making this unit one of the most versatile scanners on the market — is also available.

The SX-400 has a 20-channel computer controlled memory. Sensitivity is 0.5 μV FM/1 μV AM in the 26-300-MHz range and 0.5 μV FM/2 μV AM in the 300-520-MHz range. The unit has both a fast (8 characters per second) and slow (4 characters per second) scan rate and has a variable 0 to 4-second) delay rate. An automatic noise limiter has been added to reduce pulse noise interference on AM and FM if a filter is used to facilitate the reception of TV and FM broadcast services. For high fidelity reception of signals, channels can be spaced in either 5 or 6.25 kHz steps on VHF and 10 or 12.5 kHz steps on UHF.

An easy-to-use keypad is included for programming frequencies. The unit weighs approximately 7-3/4 pounds (3.5 kg), measures 11.8 × 3.5 × 8.3 inches (30 × 9 × 21 cm) and runs on 13.8 VDC.

The SX-400 is directly interfaceable through the RC-4000 data interface to the NEC-8801A computer. This capability allows both high speed reprogramming of the scanner's channels and an almost unlimited number of channels to be contained in the computer's memory. The computer also provides a complete record of which frequencies were received and the time at which they were received.

Other accessories for the SX-400 include an 800-1400 MHz downconverter for AM/FM (RF-8014); a 500-800 MHz downconverter for AM/FM (RF-5080); a 100 kHz-30 MHz upconverter for AM/SSB/CW (RF-1030); RF attenuators, AF gain control, Delta tuning, IF noise blanker, provision for three external antennas, and squelch. When using the RF-1030 in conjunction with the other converters CW/SSB and AM/FM in the 100 kHz-1.4 GHz range can be received. The suggested retail price is $574.90.

For more information on the SX-400 and its accessories, contact J.I.L. Industries, 17120 Edwards Road, Cerritos, California 90701.

Circle #303 on Reader Service Card.

Butternut Electronics HF2V

Butternut Electronics has just released a new two-band vertical antenna, the HF2V. Designed for 80 and 40 meters, the HF2V can be modified to operate on 160, 30, and 20 meters. The overall height is 32 feet (9.75 meters) and the antenna weighs 13 pounds (5.9 kg). The HF2V is designed to match 50-ohm cable and will give a 2:1 or less VSWR bandwidth of approximately 65 kHz on 80/75 and full-band coverage on 40 meters. Its power rating is 2 kW PEP/1 kW CW (slightly less when using the 160-meter base resonator). The HF2V is designed to operate as a 1/4-wave vertical on 40 meters and a loaded antenna on 80 meters.

For more information, contact Butternut Electronics, 405 E. Market Street, Lockhart, Texas 78644.

NEW products

automatic gain control (AGC-4)

The AGC-4 automatic gain control kit from Barrett Electronics is for any Amateur who needs to keep audio levels constant. In addition to the obvious repeater, autopatch, phone patch, or HF transmitter applications, the AGC-4 can be used in RTTY or SSTV reception with minor component adjustments. As featured in the September, 1984, issue of *ham radio*, the AGC-4 will hold a constant output of 2 volts P - P (±2 dB) with any input level ranging from −36 dBm to +10 dBm. At the center of the AGC range, the total harmonic distortion is 0.5 percent or less with a frequency response (−3 dB) that spans 40 Hz to 20 kHz. With the input shorted, the noise floor is −42 dB below the AGC output level. The AGC-4 was originally designed for the Collins/Autogram "IC" series broadcast mixer and can be operated single ended or with user supplied level or impedance matching transformers. The kit includes a drilled 2-3/4 × 1-5/8-inch printed circuit board, parts, and instructions. The unit is priced at $28.00 including U.S. shipping.

For details, contact Barrett Electronics, 525 North 2150 West, West Point, Utah 84015.

Circle #316 on Reader Service Card.

surveillance receivers

A new generation of HF surveillance receivers has been introduced by Cubic Communications with the R-3030, which features two completely independent receivers in the same rack mount normally required for a single unit.

During the first nine months of introduction, the U. S. Navy contracted for nearly 1000 receivers in the series. Principal selling points included size, light weight, and a wide range of standard advanced features.

The basic receiver is fully functional without costly options, although customer-specified enhancements such as a special data bus or up to six selectable bandwidths also are available.

The advanced modular design (in which plug-in modules are secured by 1/4-turn fasteners and can be replaced in 30 seconds or less) simplifies maintenance and improves operation. In addition, each module is independently shielded to protect circuits from electromagnetic interference as well as potential handling and storage damage.

In shock tests conducted for the Navy, the R-3030 demonstrated its rugged construction by passing a full 2000-foot-pound "hammer test" with a level A rating. The receiver was bolted to a platform, which then was struck by a 400-pound hammer swinging in successive arcs of one, three and five feet. The test was then repeated with the R-3030 turned 90 degrees.

The Navy's "level A" rating means the unit not only retained all components in place, but also continued to work perfectly after the shocks, which are meant to simulate the force of a direct hit to the ship by a non-nuclear torpedo. The reliability level resulting from this rugged construction means more than 5000 hours mean time between failures.

The R-3030's expanded performance features include: tuning in less than 8 milliseconds over a full 5 kHz-30 MHz range, 30 percent fewer parts than conventional radios and capability for full computer control. The modular, building-block approach makes it easier to meet special requirements for data bus connectors, bandwidths, AGC settings and so on.

Fault isolation is both simple and comprehensive, providing quick detection and easy replacement. All faults are reported automatically in three simultaneous modes: data bus, front panel annunciator, and LED on the faulty module. Field repair requires only removing the specified independently shielded module and plugging in a replacement. Each module is labeled and marked with a diagonal coding stripe to prevent improper installation. No special tools, alignment or adjustment are required.

The R-3030 also provides such standard features as: 100 memory channels, IEEE-488 or RS-232 bus connector modules, EMI/EMC shielding, five bandwidths (0.5-8 kHz), five operating modes (LSB, USB, AM, FM, and CW) versatile sweep and scan modes and minimal power input (approximately 35 watts per receiver).

The 48-pound dual system in a compact 5-1/4 × 19 inch rack chassis saves valuable operations and parts storage space in the field while offering effective coverage of virtually any general purpose or surveillance requirement.

For more information, contact Cubic Communications, 305 Airport Road, Oceanside, California 92054-1297.

Circle #306 on Reader Service Card.

transfer lettering fixative

Once a curiosity, dry transfer lettering is now part of the design engineer's and draftsman's tool kit. Even though the letters stick to virtually all surfaces, permanence can be a problem. Unprotected transfer lettering wears off with use. General purpose aerosol sprays aren't much help, because they contain aggressive solvents that cause dry transfers to dissolve, wrinkle or even float out of position before repeat coats can be applied to build up a protective film.

Two products from DATAK can help provide greater permanence: Dakakoat™ acrylic spray can be used to protect transfers applied to

painted or unpainted surfaces on metal and plastics. It will not, however, form a smooth film on porous materials such as paper and tracing vellum; Datakoat rests on top of the transfer, but will sink into the surrounding areas of porous materials.

The transfer lettering on porous surfces can be protected with Hardkoat,™ a unique spray for use on rubdown transfers applied to paper and other porous surfaces. A single coat softens and penetrates the transfer ink, then glues it to the surface. No additional coats are needed unless severe weathering is expected.

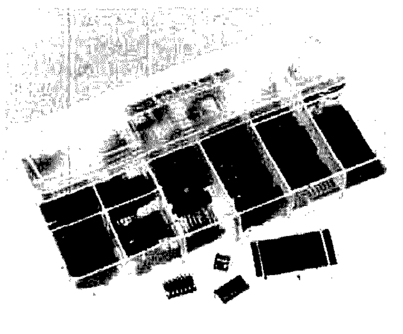
Hardkoat is supplied in 12-ounce spray cans in either gloss or matte finishes. The price for either is $4.75.

For further information, contact The DATAK Corporation, 65 71st Street, Guttenberg, New Jersey 07093.

Circle #307 on Reader Service Card.

engineer's sample case

Aries Electronics, Inc. has made available a Component Engineer's sample case, Part No. SB-100, containing over 100 pieces of various connector products the company makes. Included are sockets, Vertisockets,® elevator



sockets and single-row sockets (both stamped and collet pin versions), headers, programmable headers, switches, shorting plugs, jumper assemblies, etc. Worth over $100 if purchased individually, these parts come in a handy plastic case and sell for $30.00.

For additional information, contact Aries Electronics, Inc., P.O. Box 130, Frenchtown, New Jersey 08825.

Circle #308 on Reader Service Card.

DTMF encoder mike

Midland LMR has introduced an optional DTMF encoder microphone for its Midland SYN-TECH™ line of 2-way FM mobile radios. Available with or without automatic number identifier (ANI) capabilities, the new dynamic amplified microphones incorporate an integral DTMF encoder capable of generating the 16 standard DTMF digits 0-9 plus *, #, A, B, C and D. The

NEW products

company had previously announced availability of a DTMF decode option for its SYN-TECH radios.

The new SYN-TECH DTMF encoder microphones have a front-mounted keypad which can be enabled and disabled by an on-off switch, with current status indicated by a red LED. An internal annunciator provides audible confirmation of each keypad switch closure. Single tones can be generated, also, by simultaneously pressing two keys in the same column or row. Top-located up/down switches allow direct control of channel selection from the microphone.

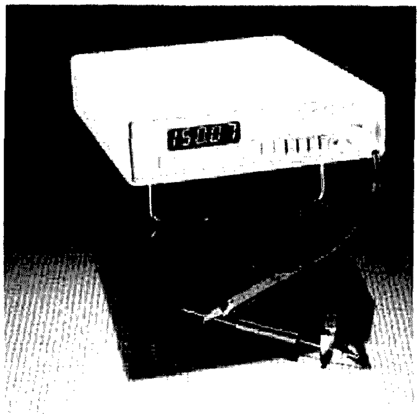
The SYN-TECH DTMF microphone with ANI capability features an ANI output which is jumper selectable from one to eight digits. The ANI sequence is automatically sent at the beginning of a transmission if a pre-set time interval has elapsed since the previous transmission. The ANI code sequence can be activated at any time by pressing the * or # key, and can be "strapped" to give single or multiple sequences when activated.

For more information on the new Midland SYN-TECH DTMF options, contact Midland LMR, Marketing Department, 1690 N. Topping, Kansas City, Missouri 64120.

Circle #309 on Reader Service Card.

micro-ohmmeter

Cambridge Technology, Inc. has introduced the Model 510, a low-cost, 4-1/2 digit, micro-ohmmeter designed to measure the resistances of switch and relay contacts, transformer and motor windings, connectors, or any other low



resistance devices. It has five ranges from 19.999 milliohms to 199.99 ohms, full-scale, 1 micro-ohm resolution, and a basic accuracy of 0.02 percent.

Three measurement modes are provided. The continuous DC mode is useful for making measurements on inductive components and the switched DC mode removes the effect of thermal voltages, the largest source of error in low resistance measurements. A pulsed mode is provided for thermally sensitive devices such as fuses. The standard unit comes with 4-terminal Kelvin test clips and a parallel BCD interface. Optional limits comparators, battery operation, and an RS-232 interface will be available in the first quarter of 1985.

For further information, contact Cambridge Technology, Inc., 2464 Massachusetts Avenue, Cambridge, Massachusetts 02140.

Circle #311 on Reader Service Card.

automatic phone patch system

NCG Co. has just announced its new Hotline 007 simplex autopatch unit. When connected between your radio and the telephone line, it enables initiation and reception of telephone calls without operator assistance. The Hotline 007 uses a unique method of signal processing that eliminates annoying squelch tails and chirps that are heard on other units. The unit uses a field programmable five digit access code to eliminate unauthorized use. In the event that you drive out of communications range, the Hotline 007 has a variable 3 to 12 minute time-out timer. When someone is calling, the Hotline 007 will page you with a CW message. To answer the call, simply send your 5-digit access code and the Hotline 007 will connect you to the phone call. You can also program the unit to refuse any calls that start with a 0 or 1. NCG also has DTMF microphones and telephone handsets available as options.

For further information and prices, contact NCG Co., 1275 N. Grove, Anaheim, California 92806.

Circle #301 on Reader Service Card.

THE GUERRI REPORT

Ernie Guerri W6MGI

toward "softer" software

Early electronic data processing systems required that people communicate with them in their own language — *machine* language. But it soon became clear that communication between people and machines in binary, or some related code, was cumbersome — for the *people* — and "higher-level" languages had to be devised that would be more recognizable to people and would automatically do conversion to machine language. FORTRAN, COBOL, BASIC, *Ada*, and LISP are all the products of a relentless move toward computer languages that are closer to human experience, yet still accommodate more complex machine architectures. This trend toward "softer" software is aimed primarily at user friendliness — the keep-it-simple concept. But there's a price to pay. Not everything is simple. In an attempt to simplify the languages, software architects have to choose the complexity of the functions they want to implement with relatively simple program statements. In this *simplification process* much computational power can be lost.

Computer architects and software designers must now work hand-in-hand to assure that commands given by humans can be interpreted by machines whose logic has been optimally structured to perform specific algorithmic functions. As machines become more complex, finding the best combination of machine structure and software language is more difficult. If we structure the machines to easily interpret "plain English," then the machines may be limited by what can be said in English. A compromise between language simplicity and machine functionality is in order. The best compromise has yet to be determined.

RF sonar serves medical needs

During 1985 some of the key biological functions of astronauts aboard the U.S. Space Shuttle will be monitored by a process called *echography*. In this process, body functions are observed by radio frequency sonar operating in the 3 to 5 MHz region. A small antenna probe is used to illuminate the tissue at very low power, and the echo returns form a time-sequenced image of the tissue. The data is stored as real-time video, and can then be digitized and image-enhanced. Resolution is very good; images are formed at rates up to 50 per second, and time-motion analysis can be updated as often as 5000 times per second. This can give very detailed information about heart movement and the flow of blood through vital organs. (You fellows with 75-meter beams, please — no elevation rotators!)

synthetic rocks

Researchers in Australia have developed a synthetic rock material that may have important application in the disposal of radioactive material. Major problems involved in disposing of high-level radioactive waste include site safety, disintegration of surrounding material, and the ability of the waste containers to remain intact for the required 10,000+ years. It's not too difficult to find geographical areas which can be extrapolated to remain stable for several hundred thousand years, but finding containers whose atomic structure is not destroyed by years of intense radiation has been an elusive process.

The newly developed "Synroc," as its developers call it, is made of three natural minerals whose atomic structure is such that high-level radiation converts the rock to a glassy material. The glassy material is atomically stable and resists cracking and subsequent leakage. The Los Alamos National laboratory is now evaluating the material by embedding high-level waste with a short half-life (less than a year) into the Synroc and then extrapolating the results. Considering the very long half-life of the more common waste, the validity of extrapolation will have to be shown. In any case, this may be a major development in what has otherwise *been an* unyielding problem.

GaAs high speed ICs make progress

Silicon semiconductor material has been the mainstay of the modern electronics industry for about 25 years. It has been continually improved and now offers good speed, high yields, and best of all, low cost. However, we are now pushing the limits of silicon physics and device geometry to get further *improvements. Gallium* Arsenide (GaAs) and an associated family of semiconductor materials

offer the higher speeds made possible by high electron mobilities. Microwave GaAs devices have been available for some time, but the price has been fairly high due to the higher material cost and poor yields. Obtaining flat, defect-free epitaxial wafers for the fabrication of GaAs ICs has been a very difficult and costly process.

However, new techniques permit material growth and processing with sufficient quality to permit small-scale integration on a commercial basis. Simple digital circuits containing tens of devices are available in the market, and arrays with a few hundred elements will be readily available during 1985. Work is progressing on very high-speed memories with cycle times of 500 picoseconds, and complex functional circuits with 1-GHz digital and 5-GHz analog circuits sharing the same real estate should be available in another year or two.

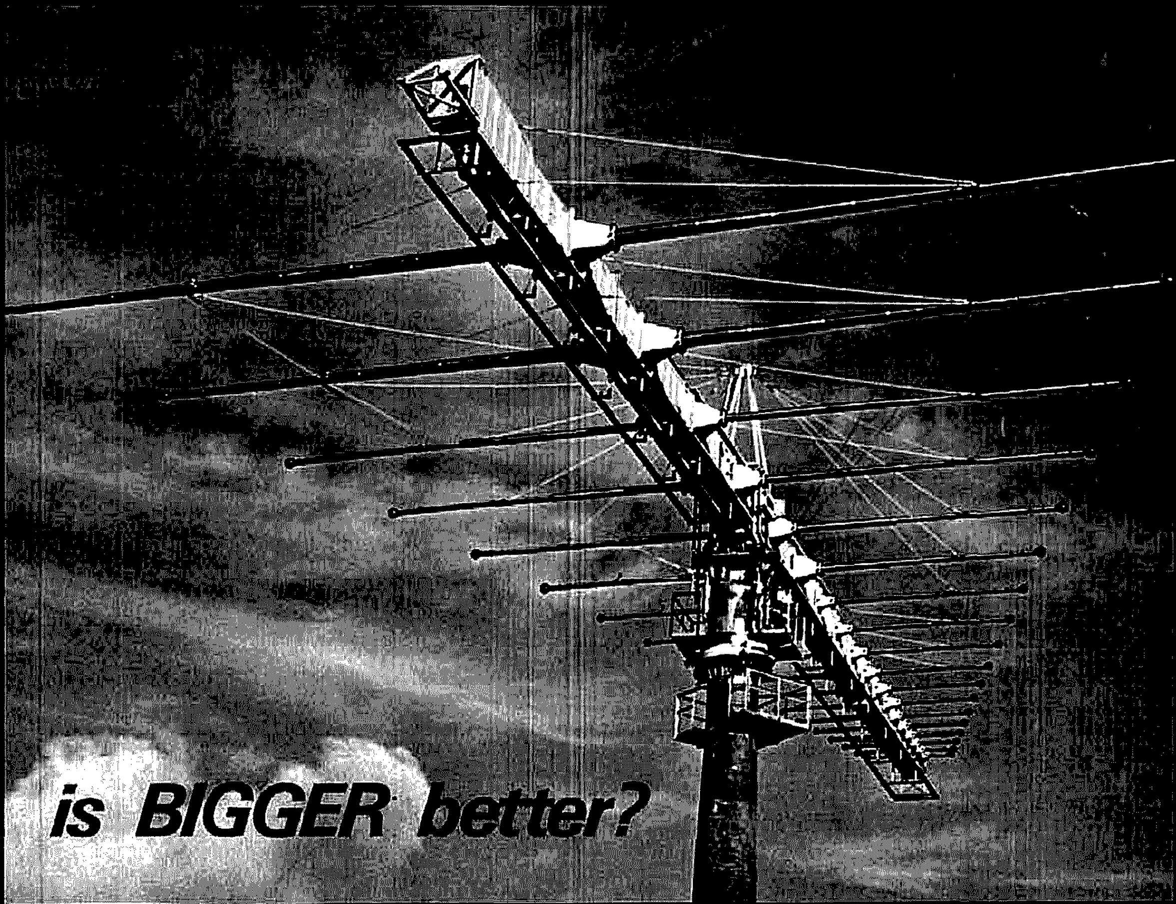
This next generation of circuit developments will offer yet another milestone in speed and capability in practically every domain of telecommunications. Amateurs have already seen the benefits of GaAs devices in low-noise VHF/UHF amplifiers, and will next see comparable performance in the signal processing sections of new radio equipment.

ham radio

MAY 1985 / $2.50

ANNUAL ISSUE

ham radio magazine



in this issue: stacking Yagis properly • 0.5-30 MHz active antenna • vary your take-off angle • switchable vertical wire array • alternative method of feeding phased arrays • multiband sloping vee beam • 160-meter transmission line antenna • RF transmission cable for microwave applications • baluns • convert your fixed tower to a tilt-over • plus W1JR, W6SAI, K0RYW, and the Guerri report

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contents

REFLECTIONS

the readers speak . . .

We asked — and **you** answered our call for solutions to the problems facing our hobby today (see "Reflections," February, 1985, page 5. Each and every one of your letters was read, and we're proud to act as a clearinghouse for your many ideas and suggestions.

Foremost you recognized that Amateur Radio has changed, and that hams have to accept that fact before any effective improvements can be made.

Here are some of your comments on the problems, followed by some suggested solutions:

"The challenge is gone. Want to hear China? Just turn on the TV."

"Equipment is just too expensive and complex for us to do our own repairs."

"Our kids are into computers, not radio."

"Hams are boring and don't know how to communicate any more."

"Consumer electronics and RFI — what a nightmare!"

"Enforcement, what a joke — the FCC's a paper tiger."

Can we expect the general public to help? I doubt it. Many of them see us as "the ham down the block with the ugly tower and wires all over." Don't expect too much sympathy from the neighbors — especially if their brand-new VCR is carrying our conversations as well as their programming. So what do we do?

SCHOOLS. Talk to your local school board. Could some space and time — say an hour a week — be set aside for a Novice course taught by local hams? What about high school science teachers? Would they be interested in having guest speakers address classes on Amateur Radio subjects? Radio clubs, how about sponsoring radio-related science fairs, with prizes or scholarships awarded for the best projects? How about donating software — morse code tutors and technical Q&A's, for example — to school libraries? Besides the school environment, summer camps and community centers would also be logical places for reaching and teaching prospective hams.

PUBLICITY. We need more of it — the good kind, that is. Make a personal effort to contact and provide details of Amateur Radio events to your local newspaper and radio or TV stations. Every little bit helps. There's a move afoot to produce and distribute a brochure describing the "fun" aspects of our hobby. When it becomes available, get some and hand them out to the neighborhood kids. Do any of you work in the media or public relations? How about using your skills to help improve our image?

PARTICIPATION. This means all of us. No contribution is too small. Participate by helping out at the club, helping newcomers to the hobby even *after* they're licensed. (This in itself might help reverse the "dropout" rate.) How about being an Elmer and forming a "ground wave net" to bring together new hams in a local net designed to help them overcome their initial shyness? I'm sure we can all remember how we felt during those first few weeks on the air.

Dear Senior Citizens, you have the time, the knowledge, and the political clout. Your active leadership and participation are essential.

What would happen if, for the good of Amateur Radio, we set aside our individual concerns and pooled our talents in a single "umbrella" organization dedicated solely to protecting and preserving our hobby? Perhaps then it would be more difficult for commercial interests to nibble away at our spectrum. We have to lobby from a position of unity and strength.

PERSONAL GROWTH. Hams used to be interesting people to talk to. We could converse on diversified subjects at length. Now too many of us appear to be overly involved in contesting and card-collecting — i.e., doing the same old thing all the time, when we should be advancing our knowledge, experimenting with different modes, and searching for new frontiers.

Perhaps we can act on some of the advice offered by our readers and help improve the hobby.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

THE 2-METER CHANNEL SPACING ISSUE MAY RESULT IN A CONFRONTATION that could leave Amateur Radio with some bad long-term wounds. Repeater operators from the mid-Atlantic and New England states met March 2 to strongly endorse 15 kHz spacing not only for 146-148 MHz, but even possibly for 144.5-145.5 MHz as well, thus picking up another six channels. In a similar action southern California's 2-Meter Area Spectrum Management Association met March 23 to also endorse 15 kHz spacing across the whole band, despite serious interest in northern California in shifting to 20 kHz as their neighbors in Oregon and Washington have done. By their action the southern California group's decision puts them in direct conflict with Mexican Amateurs, whose government has mandated 20 kHz on the 2-meter band. It's believed that Mexican representatives who attended that meeting may even ask their government to lodge a formal objection with the U.S. State Department.

20 kHz Proponents Seem Equally Determined In Their Choice, with Texas the latest in a number of western states to endorse the change. Michigan was the first state east of the Mississippi to adopt 20 kHz, and it's under heavy pressure to reconsider. Alabama is reported to have voted to implement 20 kHz by July 1, but Missouri opted to stay at 15 kHz at a February meeting. Midwestern repeater representatives plan to meet at Dayton for further discussions.

Despite The FCC's Strong Desire For A National Coordinator, formalized in PR Docket 85-22, the escalating 15 vs 20 kHz conflict seems to offer little hope for simple or early resolution of the increasing number of repeater conflicts that led to that NPRM. Indeed, the FCC may even be asked to endorse 15 kHz through a Petition for Rule Making. Such a move was considered by the East Coast group at its recent meeting in support of 15 kHz. However, it is hard to see how such a petition would find any support at all within the Commission, which is trying very hard to get away from regulating details of Amateur operations. Furthermore, in this case such an FCC mandate would put it in direct conflict with the Mexican government and thus risk diplomatic repercussions.

AMATEUR RADIO PROBABLY WON'T FLY ON THIS SUMMER'S SPACE SHUTTLE, according to late word from NASA. The official reason is problems of integrating the exotic Amateur equipment that Tony England, WØORE, was to use on Flight 51F with shuttle equipment. Though strong last-ditch efforts are being made to salvage the Amateur operation, at presstime it looks very much like there won't be an Amateur operating from space this year.

AMATEUR RADIO VS CABLE TV IS YET ANOTHER CONFRONTATION that seems to be escalating. In late February the FCC proposed in Mass Media Docket 85-38 that cable TV radiation limits be relaxed by another 8 dB, to 50 microvolts per meter, and that various other restrictions including the requirement for annual system inspections be dropped. In view of the number of Amateurs who've had problems with cable leakage interfering with their 144 or 220 MHz operations, this proposal presents a real threat to Amateur Radio. Comments were due in late March. Amateurs have a strong ally against the proposal in the broadcast industry.

Amateurs Have Been Shut Down Due To Cable Interference in two recent unrelated cases. In the first, WB2OTK/4 was ordered off the air by the Engineer-In-Charge of the FCC Atlanta Field Office for interference with the Greenville, S.C., cable system. FCC Amateur rule 97.73(d) on interference to other stations was cited, despite checks that showed the cable system was so badly maintained that a 150 mW hand-held's signal caused problems for viewers. The FCC inspected WB2OTK's station but did not even contact the cable operator in an apparent violation of the FCC's own rules requiring a cable system to take responsibility for resolving interference problems. In the other case, WB4NMA was shut down, again by the Atlanta Field Office, for interfering with cable viewers in Gainesville, Georgia.

2-Meter Privileges Were Restored To Both Operators following the intervention of the ARRL through its General Counsel Chris Imlay, N3AKD. However, the Commission's proposed cable rules relaxation combined with the kind of FCC Field Office attitude that led to these two cases could spell serious trouble for future Amateur Radio VHF/UHF operation.

BASH EDUCATIONAL SERVICES IS OUT OF BUSINESS. Writing to subscribers to his FCC Rules Part 97 update service, Dick Bash, KL7IHP, stated that his FCC exam study materials business had subsidized the rules updating service but, under the volunteer exam program, there was no longer a unique niche in training for him to fill. Bash was very controversial for his publishing of word-for-word correct answers to all FCC-administered Amateur exams.

A SUGGESTION THAT ARRL ASSUME CALLSIGN ASSIGNMENT RESPONSIBILITY is receiving some interest in Washington, as the FCC looks at tighter budgets. It's likely, however, that callsign responsibility would be delegated only as part of a total program that would include the entire licensing function now done by FCC's Gettysburg facility.

THE ANTENNAS VS LOCAL RESTRICTIONS ISSUE seems to be heading toward an FCC Notice of Proposed Rule Making, which will probably consider it from the private satellite dish viewpoint. Just what this means for PRB-1 and Amateur Radio remains to be seen.

NEW WARC BAND AVAILABILITY HAS BEEN DELAYED by at least three months. An extension of the Reply Comment period, requested by the Personal Radio Steering Group, bumped expected action on the 10, 24, and 902 MHz bands off the FCC's first quarter agenda. The PRSG is believed to be pushing for allocation of part of 902-928 MHz to a Personal Radio Service.

A NATIONAL VEC NET IS OPERATING on 20 meters Sundays, starting at 1700Z. The 40-meter midwestern VEC net has moved to 1800Z, still on 7280 kHz. DeVry is now officially a national VEC.

stacking Yagis
is a science

Determining
E/H plane patterns
results in
optimized performance

Several years ago I began a project designed to develop a means of determining gain of a particular 144 MHz Yagi. Knowing that number, I would be able to calculate the aperture. Once the aperture was known, I would know the correct stacking distance and would then be able to build the perfect 144-MHz array.

As one might expect, things didn't quite work out as planned. But eleven years and seven EME arrays later, I believe I have found an improved method of determining the optimum stacking distances for multiple Yagi VHF and UHF arrays.

what is optimum stacking?

"Best array performance" is a matter of opinion, and depends on how the antenna is used; an array can be optimally stacked, for example, to deliver maximum gain, or for a controlled azimuthal pattern that produces a deep null in the direction of interference. This article primarily addresses Yagi stacking for use in EME (Earth Moon Earth or moonbounce) communications. However, this technique is also appropriate for other space communications applications such as satellite communications and radio astronomy. It could also be used for land-based communication requiring high gain such as tropospheric scatter.

For EME and other weak-signal VHF/UHF work, optimum stacking distance can be defined as that distance which yields the greatest array gain versus lowest array temperature. Used by professional space communications engineers, this definition refers to maximizing the G/T ratio. With Yagis, the best G/T usually never occurs at the distance which yields maximum stacking gain. It normally occurs at significantly closer spacing. In simpler terms, optimum stacking distance is that stacking distance which yields the greatest array gain increase while simultaneously keeping all sidelobes at an acceptable level.

the effect of antenna temperature

Several good sources of information are available for those unfamiliar with the concept of antenna temperature.[1,2] To review, antenna temperature is the temperature of the object at which the main lobe of the antenna is pointed — i.e. the Earth, Moon, Sun, hot sky, or cold sky. However, if the array has sidelobes of significant area and amplitude, unwanted noise can be picked up from noise sources in the direction in which the sidelobes are pointed. This unwanted noise reception will increase the net antenna temperature. (**Table 1** lists typical temperatures of several "objects" at 144 MHz and 432 MHz.) A 432-MHz array with large sidelobes pointed at the Earth will suffer a significant receive signal-to-noise loss because of the reception of Earth noise; likewise, a 144-MHz array with sidelobes directed toward a hot sky or a leaky power line will experience a similar degradation in receive performance.

table 1. Typical temperatures of "objects" at 144 MHz and 432 MHz.

object	144 MHz degrees K	432 MHz degrees K
cold sky	175	10
hot sky	3,200	190
earth	290	290
arcing power line	100,000	6,000

By **Steve Powlishen, K1FO,** 816 Summer Hill Road, Madison, Connecticut 06443

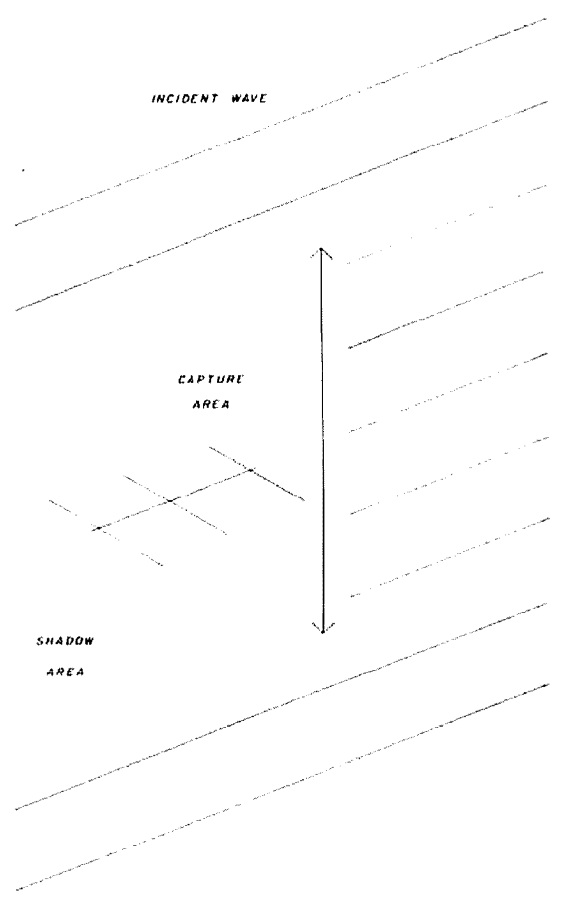


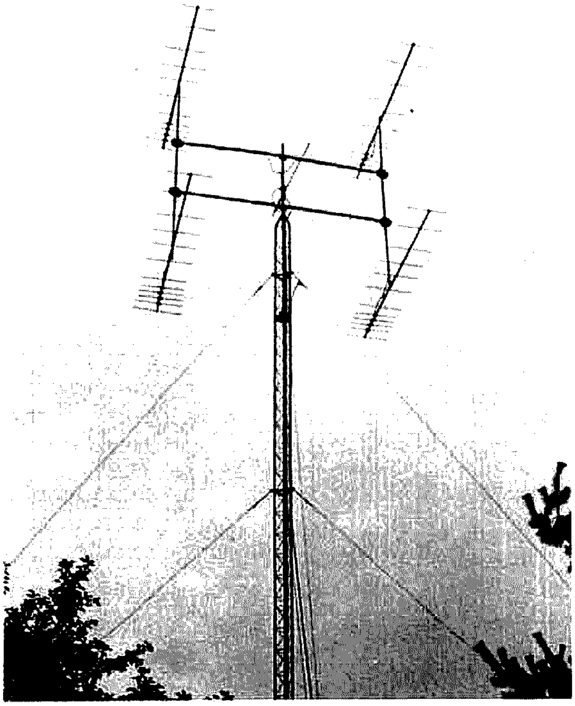
fig. 1. Antenna aperture and shadowing effect.

As a point of reference, commercial satellite Earth station antennas that use parabolic dish antennas with Cassegrain or Gregorian subreflector systems have antenna temperatures as low as 18 degrees K. In comparison, an Amateur dish using a simple dipole or horn feed at 432 or 1296 MHz has a typical antenna temperature of 65 degrees K, while a 16-Yagi 432-MHz array with the Yagis spaced for maximum gain may have an array temperature over 170 degrees K (or even worse if low-loss phasing lines are not used). Conversely, stacked 432-MHz EME Yagi arrays incorporating optimized stacking distances have been built to provide array temperatures lower than 85 degrees K — including phasing line losses. These antenna temperatures were measured by pointing the antenna at the Earth and then at the cold sky and comparing the noise ratios. The noise contribution due to the receiver can then be factored out if the receiver system noise temperature is accurately known. The significance of the lower array temperature cannot be overstated. If we assume a high performance receiver with a 0.45 dB system noise figure, lowering a 432 MHz array temperature from 170 degrees K to 85 degrees K results in a receive signal-to-noise improvement of about 2.3 dB or almost the equivalent of doubling the array size! These results have been obtained using standard Yagi designs such as the NBS Yagis and the K2RIW 19-element Yagi. I expect further improvements can be obtained when Yagis designed with best G/T in mind are available to Amateurs.

methods of determining optimum stacking distances

There are three methods of determining optimum stacking distances. The first method to be examined briefly, is based on classic antenna theory. The second, which will be emphasized, is experimental. Computer analysis, currently used by the professional community is the third, and will not be discussed here. With programs for Yagi analysis now readily available to the Amateur, it is hoped that the more mathematically inclined and computer-knowledgeable Amateurs will carry on where this article leaves off and extend computer modeling to include optimum stacking.

The concept of antenna directivity, (**fig. 1**), put forth by Kraus[3] and introduced to Amateurs by Orr and Johnson,[4] holds that all antennas have an effective capture area, or area around the antenna that "captures" or extracts the electromagnetic energy from space. The higher the gain of the antenna, the


Four 12-element LPYs stacked on a telescoping H-frame.

larger the area from which energy will be extracted. Behind the antenna there will be a shadow area, or space where the field strength of the incident wave is reduced in magnitude. (This concept is analogous to putting an object in front of a light source and creating a shadow behind it.) In mathematical terms the capture area is directly proportional to gain and is defined in **eq. 1**.

$$A_{em} = 0.13 \cdot 10^{dBd/10} \qquad (1)$$

Where A_{em} is the effective capture area in wavelengths squared and dBd is the gain of the antenna in decibels over a half wave dipole. For antennas such as Yagis, which have an elliptically shaped aperture, the size of the effective aperture will be slightly different between the E and H planes. The aperture dimensions in wavelengths squared can be calculated by using **eqs. 2** and **3**.

$$A_H = 2\sqrt{\frac{A_{em} \cdot \theta_E}{\pi \cdot \theta_H}} \qquad (2)$$

$$A_E = 2\sqrt{\frac{A_{em} \cdot \theta_H}{\pi \cdot \theta_E}} \qquad (3)$$

Where A_E is the E-plane aperture dimension, A_H is the H-plane aperture dimension, θ_E if the E-plane half power beamwidth, and θ_H is the H-plane half power beamwidth.

There are two problems with using these formulas to calculate stacking distances. First, an antenna's aperture is not an ellipse with a clearly defined boundary, with radio waves being extracted on one side of the boundary and nothing happening on the other side. Instead, an antenna progressively extracts less and less energy from space continuously. In addition, the half power beamwidths of an antenna are merely a point on the field strength gradient of the antenna. Therefore, proper stacking distance becomes a question of determining where two *unclearly* defined volumes separate. It is not a solid boundary like a brick wall.

The second problem, largely self-inflicted by Amateurs eager to believe they could defy the laws of physics and discover something that antenna engineers could not, is that of believing inflated gain figures produced by both Amateurs and some manufacturers of Amateur antennas. (In defense of Amateur equipment manufacturers, their claims are restrained in comparison to their CB and home TV counterparts). Using these optimistic gain claims, which in some cases are typically 3-dB high, leads to arrays which have grossly oversized stacking dimensions. The gain figures shown in the recommended stacking distance

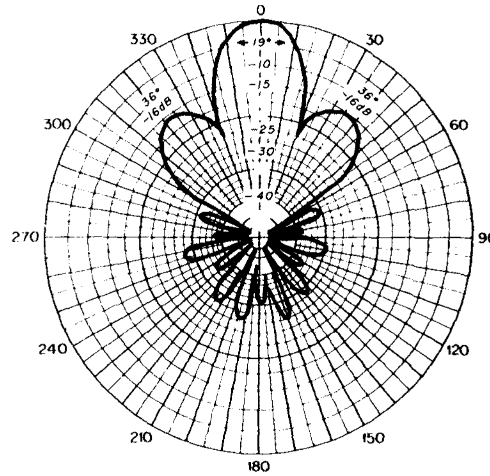


fig. 2. Four stacked 12-element Swan 2-wavelength LPY array E-plane pattern with 10-foot (1.5 wavelength) E and H-plane spacing.

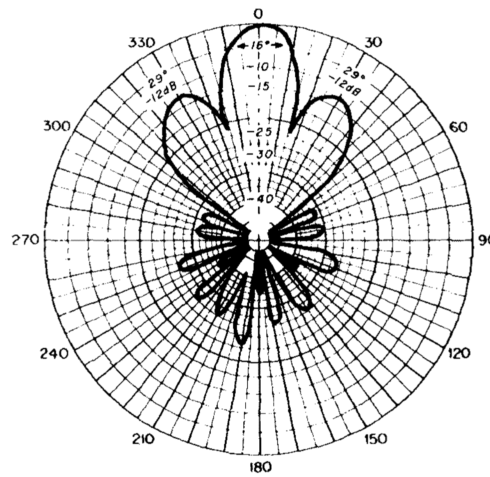


fig. 3. Four stacked 12-element Swan 2-wavelength LPY array E-plane pattern with 12 foot 0 inch (1.76 wavelength) E-plane and 11 foot 4 inch (1.66 wavelength) H-plane spacing.

table (**table 2**) represent hundreds of hours of antenna measurements I have performed, the compilation of data from ten years of antenna gain contests, and finally, computer analysis of almost all the antennas listed in the table. The result is a listing of gain figures closer to reality, I believe, than anything previously available. (For further discussion of how much gain can be expected from a given boomlength Yagi, reading DL6WU's article is suggested.)[5]

measuring antenna gain

As part of my experiments to measure antenna gain and determine proper stacking distances, a telescoping H frame was constructed on my 60-foot (18.3 meters) tower to measure the 144-MHz antenna. This arrangement allowed for rapid placement of Yagis along with a simple method of adjusting the spacing to any separation of up to 16 feet (4.9 meters). A smaller frame (6 feet/1.9 meters) was used to measure the 432-MHz Yagis.

The next step was to figure out a relatively accurate means of making pattern measurements. Over the years I tried a variety of measurement techniques, including use of strip chart recorders, spectrum analyzers, and RF voltmeters. Out of all this came a relatively simple and reasonably accurate method that is within the reach of most serious VHF/UHF experimenters.

The basic requirements for pattern measurement are:

- an accurate direction indicator;
- a receiver with a calibrated signal strength indicator; and
- a signal source located so as to minimize reflection problems.

Satisfying each requirement was surprisingly simple. The direction indicator I used was a selsyn readout* with close to 1 degree accuracy. Alternatively, the digital readout system using 10 turn pots and popularized by many EMErs[6] could be used. The availability of signal generators such as the Hewlett-Packard 608 series (or their military counterparts, the TS-510 series)* solves the receiver calibration problem. My method of measurement consisted of connecting a digital voltmeter (DVM) to the AGC (automatic gain control line) of the station receiver (an R-4C and TR-7). Care must be taken to keep signal strengths high enough above the noise (floor) to eliminate signal-to-noise plus noise ratio correction problems. The signal level must not be so high that it causes gain compression problems. The quick way to get a pattern was to run the antenna through 360 degrees while recording the AGC voltage readings on the DVM. The antenna was then replaced with the signal generator and the AGC readings were converted to a dB scale. (A 10 dB attenuator was placed between the converter and antenna or signal generator to eliminate impedance problems.) This "calibration" of the receiver was done after every measurement to eliminate receiver gain drift errors.

The signal source presented the trickiest problem, but again a simple solution was found. A number of signal sources were tried including locating signal sources in my back yard, in the woods about 1000 feet (305 meters) away, and at a local ham's QTH about 2 miles (1.3 km) away. All of these solutions gave marginally repeatable results. That is, an antenna would look different from day to day and from source to source. Because I believed the problem to be reflections from various objects, I decided to try a signal source located above the clutter. My location at that time was 105 feet (32 meters) above sea level. The antennas were located at 65 feet (20 meters), well above the nearby trees. My location was surrounded by hills up to 1200 feet (366 meters) high, 10 to 12 miles (16 to 19.2 km) away. The ground between my location and the hills dropped in elevation, which made the tops of the hills a completely clear shot at about a 0.5 degree elevation angle. A fellow ham located on one of the hills was called upon; by using high gain source antennas (stacked 3.2 wavelength NBS Yagis on 144 MHz and RIW19s on 432 MHz), repeatable pattern measurements became a reality. After performing many measurements I was able to "calibrate" my range such that I could see how different antennas gave the same false lobes or left to right unbalance. The patterns shown in this article have been cleaned up to eliminate known range errors. Various NBS antennas were constructed and their patterns measured. My measured patterns correlated very well with the NBS published patterns and offered proof of the method's validity.

It was also found that accurate relative gain measurements were possible. Gain measurement calibration was made by putting two identical antennas on the frame and measuring the relative signal level between them. The antennas were then switched in position and the measurement repeated. In this manner any differences in signal strength could be factored out before a test antenna was measured. Results were found to be repeatable to within 0.2 dB over the two-year period the gain tests were made. A similar method was used to measure the gain increases to be had from stacking antennas.

the effects of antenna spacing

To get an indication of what happens when the spacing of Yagi arrays is changed, an array of four 12-element LPY (log periodic Yagis, as introduced by Oliver Swan and later produced by KLM) was set up on the telescoping H-frame. Pattern measurements were made at one foot (0.3 meter) spacing increments from 10 feet (3.0 meters) up to 16 feet (4.9 meters) (1.4 to 2.3 wavelengths). The resultant patterns are

*Available from Fair Radio Sales, P.O. Box 1105, Lima, Ohio 45802

typical of all two-wide Yagi arrays. As the spacing between Yagis is increased, the main lobe beamwidth narrows, the sidelobes increase in amplitude and the nulls get much deeper. The E-plane pattern at 10 feet (3.0 meters), 12 feet (3.7 meters) and 14 feet 4 inches (4.4 meters) spacings is illustrated in **figs. 2**, 3, and 4. Looking at the sample patterns, it can be seen that the beamwidth of the first sidelobes is close to that of the main lobe for each spacing and that its amplitude increases rapidly at larger separations. Other minor lobes start to have significant amplitudes at the greater separations.

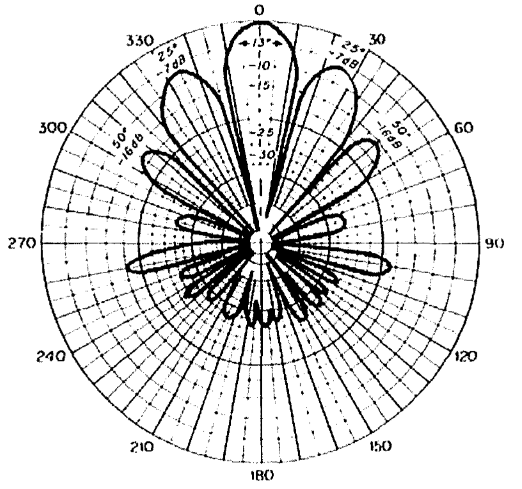


fig. 4. Four stacked 12-element Swan 2 wavelength, 144-MHz LPY array E-plane pattern with 14 foot 4 inch (2.1 wavelength) E and H-plane spacing.

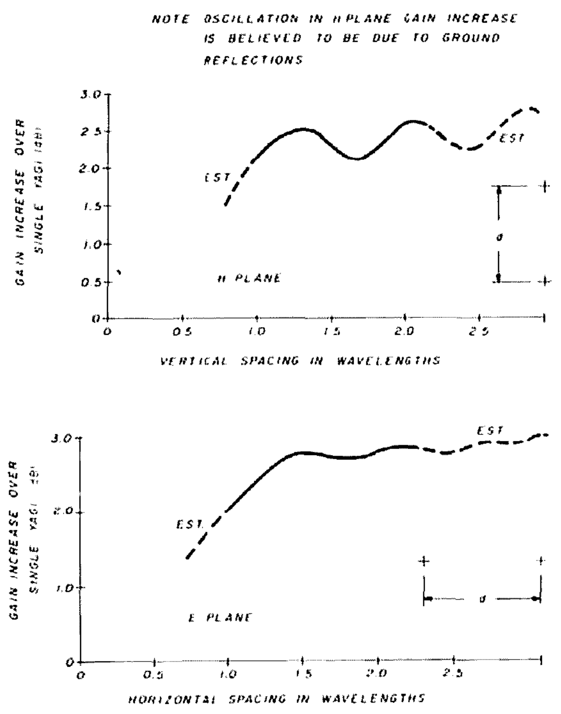
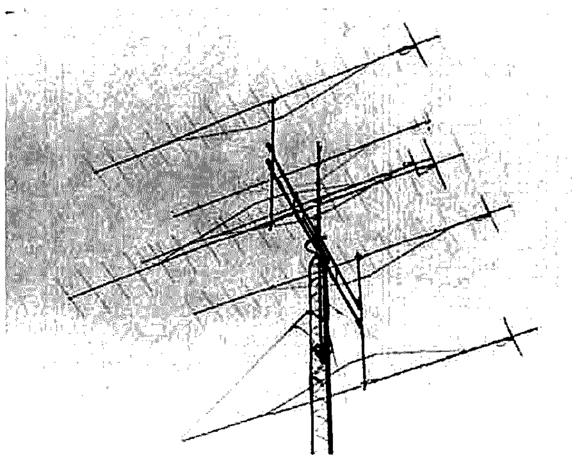


fig. 5. Array gain versus spacing using Swan 12-element, 2 wavelength 144-MHz LPYs.

The question of which distance is the proper one still remained to be answered. Gain and individual pattern measurements on the 12-element LPY indicated an actual gain of about 11.2 dB over a half-wave dipole (dBd). By using **eq. 1, 2,** and **3** (single antenna pattern of 36 degrees E-plane by 40 degrees H-plane), stacking distances of 10 feet 8 inches E-plane (3.3 meters) 9 feet 7 inches (2.9 meters) H-plane were calculated. The 10 foot 8 inch (3.3 meter) E-plane spacing gave a pattern similar to **fig. 2**, but with first sidelobes down 14 dB. Next relative gain measurements were attempted between one and two antennas. Resultant gain curves showing stacking gain versus spacing are shown in **fig. 5**. The general shape of the curves are similar to those given in the now-famous NBS Technical Note 688,[7] (**fig. 6**) except I did not see any gain decrease as the spacing was increased to large distances, only a flattening of the gain curve. I also saw an apparent larger gain increase in the E rather than the H-plane. This is again indicated in NBS Note 688. As illustrated in **fig. 5**, the knee in the gain increase occurs at about 2.7 to 2.8 dB in the E-plane and at 2.5 to 2.6 dB in the H-plane. (When phasing line losses are factored out.)

Next, I attempted to relate first sidelobe levels to position on the gain increase curve and found that the gain increase started to flatten out when the first sidelobes were −12 to −13 dB down. It should be pointed out that at the time these measurements were made,



Four 17-element, 3.2 wavelength NBS Yagis stacked on an H-frame.

very wide spacing was very popular and many Amateurs, including myself, thought the NBS report with its relatively close spacings was in error. The correlation between my measurements to the NBS curves

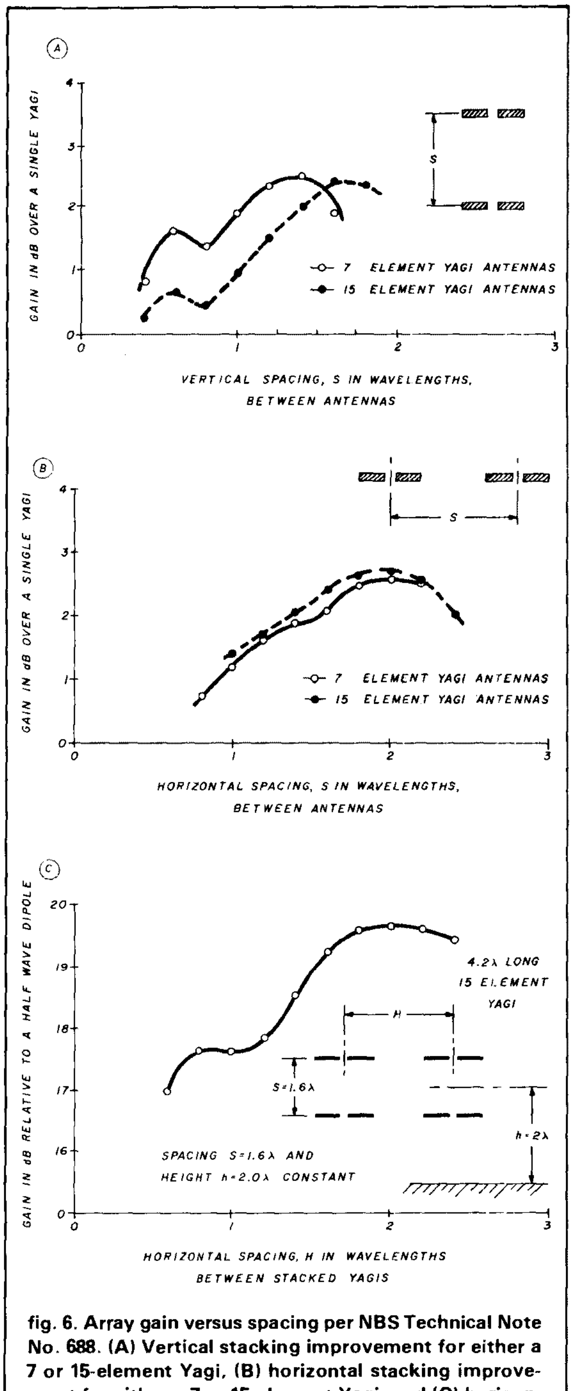


fig. 6. Array gain versus spacing per NBS Technical Note No. 688. (A) Vertical stacking improvement for either a 7 or 15-element Yagi, (B) horizontal stacking improvement for either a 7 or 15-element Yagi, and (C) horizontal stacking improvement for two sets of stacked Yagis spaced vertically 1.6 wavelength apart.

was truly amazing, especially considering that I was intent on proving NBS *wrong*.

The additional rule of thumb used by Amateurs was to stack Yagis so that the first sidelobes were 13 dB down. This seemed to correspond to where the gain increase curves flattened out; however, when the sidelobes were −13 dB, the main lobe was less than one half that of a single antenna. When I attempted to find out where the −13 dB rule of thumb came from, the only explanation I could find was that two sidelobes at −13 dB were, in total, −10 dB from the main lobe and anything 10 dB down (or 1/10 amplitude) was insignificant. This seemed plausible except that the H-plane should have a pattern similar to the E-plane — and if it also had two sidelobes 13 dB down (with similar beamwidth to the main lobe), the sum of just those four lobes would be −7 dB relative to the main lobe or 20 percent of the amplitude of the main lobe. Thus if all four sidelobes were looking at noise sources 10 times stronger (in reality, it is not very likely all the sidelobes would be facing similar noise sources) than the background noise the main lobe was looking at, the array would suffer a 6 dB signal-to-noise loss on receive.

improving a 144-MHz EME array

With this information in mind, I began to look at the 144-MHz EME array I was using at that time. It consisted of 4 Cushcraft A32-19, 19-element 3.2-wavelength 22 foot (6.7 meter) long Yagis patterned after the NBS 17-element Yagi, with a tri-reflector added. The gain of a single A32-19 is about 13.2 dBd with −3 dB beamwidths in the E and H-plane of approximately 28 by 33 degrees. By using the previously defined aperture calculation method, spacings of 2.1 by 1.75 wavelength or 14 feet (4.3 meters) E-plane by 12 feet (3.7 meters) H-plane were calculated. The manufacturer of the antenna was recommending the same spacings, so it seemed reasonable to use them when constructing the array.

The performance of the array seemed acceptable. I usually received good signal reports, but on receive, signals always were poorer than expected. I easily dismissed the lack of hearing on a noisy urban environment. Looking at the NBS stacking curves, NBS was recommending 2.0 by 1.6 wavelength spacing for the 15-element 4.2 wavelength Yagi, an antenna with about 0.8 dB more gain than the A32-19. I then made some sidelobe measurements and found that the first E-plane sidelobes were down about 12 dB and the H-plane sidelobes were down only 10 dB. A quick decision was made to move the antenna spacing in to 1.9 by 1.6 wavelenghts or 13 by 11 feet (4.0 by 3.4 meters) E by H-plane, respectively. The results were startling. During the first two months of operation at the closer spacing, about 20 new stations were worked on EME,

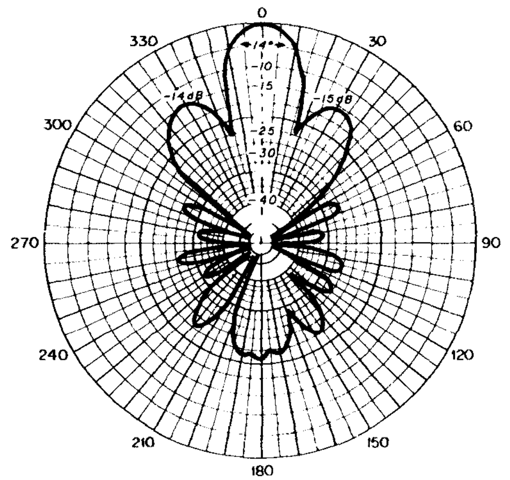


fig. 7. Four stacked NBS 17-element, 3.2 wavelength Yagi E-plane pattern with 13 foot 0 inch (1.9 wavelength) E-plane and 11 foot 0 inch (1.61 wavelength) H-plane spacing.

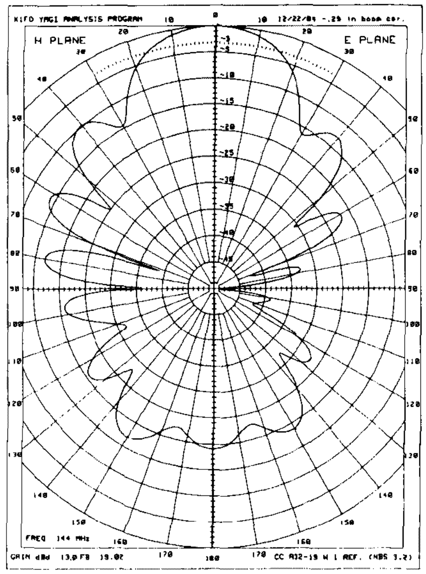


fig. 8. NBS 17-element, 3.2 wavelength Yagi computer generated E and H-plane patterns.

many of them stations I had tried to work in the past without success.

After that I obtained four more A32-19 Yagis. I modified them to standard NBS dimensions and removed the tri-reflectors. I then set them up on the telescoping H-frame and set out to find what was going on. The E-plane pattern of the 17-element 3.2λ NBS array spaced at 13 feet (4.0 meters) E-plane is shown in **fig. 7**. The pattern looks very good, with first sidelobes down 14 to 15 dB and all other lobes down 25 to 30 dB. H-plane patterns are usually more difficult to measure. Reflections from objects such as trees and utility poles, which are essentially vertically polarized, complicate the problem. Tilting the array back causes changes in ground reflections, which can induce errors if that method is used. Because of that I did not make a complete H-plane pattern measurement, but I did check the first H-plane sidelobes and found them to be only 12 dB down at the 11 foot (3.4 meter) spacing. The aperture calculations indicated that the spacing was already too close — however, NBS had indicated that 1.6-wavelength or 11-foot (3.4 meter) spacing was correct for the higher gain 15-element Yagi. To explain this wide discrepancy between calculated spacings and measured patterns, the patterns of the individual Yagis were examined. **Figure 8** is a computer-generated plot of the E and H-plane patterns of the 3.2 wavelength NBS Yagi. The H-plane pattern is noticeably less directive than the E-plane with larger sidelobes over the entire pattern. The array pattern of a number of Yagis is the resulting interference pattern of the individual Yagi patterns interacting with each other. It follows that the resulting array pattern of multiple Yagis will have larger sidelobes in the H-plane.

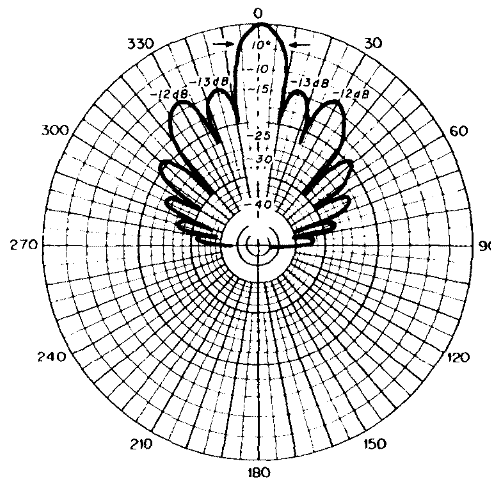


fig. 9. Six stacked NBS 17-element, 3.2 wavelength Yagi half H-plane pattern with 13 foot 0 inch (1.90 wavelength) (2 wide) E-plane and 11 foot 0 inch (1.61 wavelength), (3 wide) H-plane spacing.

While measuring the four 3.2-wavelength NBS Yagi array, it was decided to attempt to relate array main lobe beamwidth to sidelobe level and array gain increase. The 3.2 wavelength NBS Yagis reacted similarly to the 12-element LPY antennas previously meas-

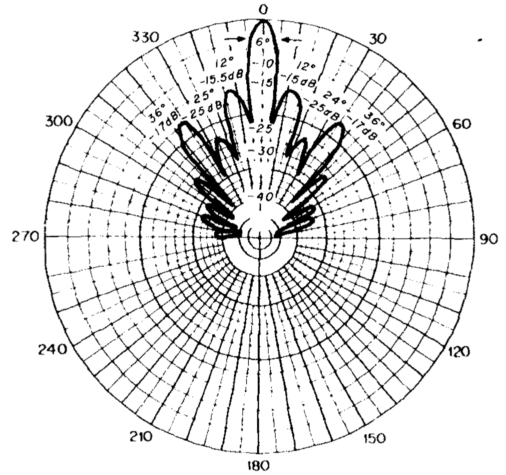


fig. 10. Sixteen stacked Cushcraft 424B Yagi E-plane pattern with 60 inch (2.2 wavelength) E-plane and 58 inch (2.1 wavelength) H-plane spacing.

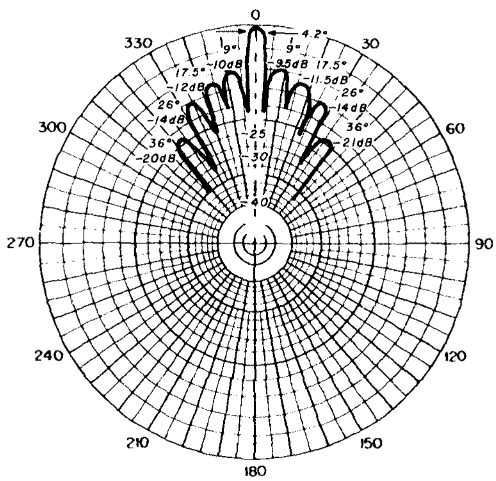


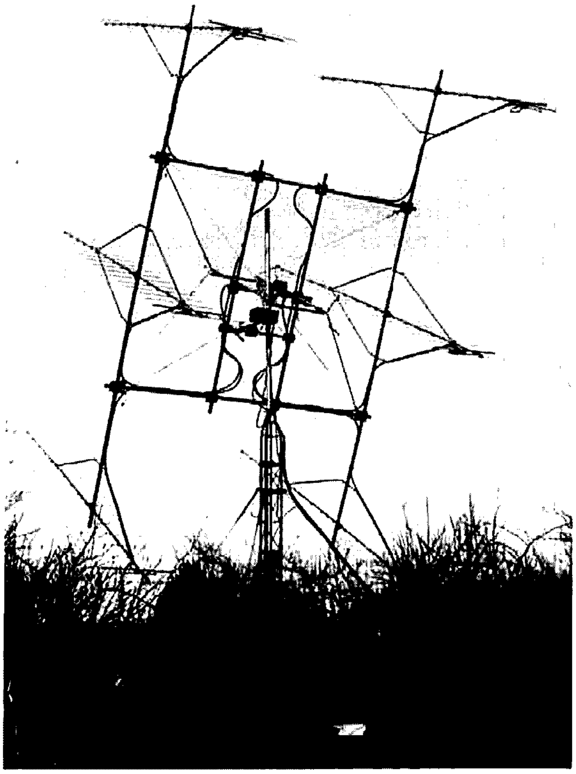
fig. 11. Sixteen stacked Cushcraft 424B Yagi H-plane pattern with 60 inch (2.2 wavelength) E-plane and 58 inch (2.1 wavelength) H-plane spacing.

ured. When the first sidelobes were −13 dB, the main lobe was narrower than one half that of a single Yagi. Likewise, the main lobe was approximately one half the beamwidth of a single Yagi when the first sidelobes were −14 to −15 dB. *This relationship of first sidelobes at −14 to −15 dB when the main lobe beamwidth is half that of a single Yagi has held up in all subsequent arrays I have measured.* This also includes arrays that are three and four Yagis wide where the array −3 dB beamwidth is approximately equal to the beamwidth of a single Yagi divided by the number of Yagis in that plane. As an example, my 144-MHz array was expanded to six 3.2 wavelength Yagis. The vertical spacing was kept at 11 feet (3.4 meters). The array's H-plane pattern is shown in **fig. 9.** The number of major sidelobes in an array is equal to the number of elements in a plane minus 1. Thus the six-Yagi array will have two major H-plane sidelobes. In this case, the first sidelobe is −13 dB down and the second is −12 dB down. As expected, the main lobe is narrower than one third that of a single Yagi at 10 degrees. Note that as the number of elements in an array are increased (and consequently the number of major sidelobes increase) it becomes much more important to keep the sidelobe amplitudes under control. A look at a 16 Yagi 432-MHz EME array will expand on this point.

more dramatic results at 432 MHz

Frank Potts, WA1RWU, had erected a 432 MHz EME array consisting of 16 Cushcraft 424B 24-element 7.6 wavelength Yagis. The instruction sheet for the 424B recommended 66-inch E-plane by 60-inch H-plane spacing. This was considerably closer than the spacings determined from calculating the aperture. Based on actual antenna gain of 15.8 dBd and a pattern of 20 degrees by 22 degrees (E by H), the spacings were calculated to be 72 inches (1.8 meters) by 65 inches (1.7 meters) E-plane by H-plane. When Dave Olean, K1WHS, of Cushcraft was contacted, he recommended the use of even closer spacings for an EME array — as close as 60 inches (1.5 meters) by 54 inches (1.4 meters). Because of mechanical considerations, the array was assembled using 60-inch (1.7 meter) horizontal (E-plane) by 58-inch (1.5 meter) vertical (H-plane) spacing. Phasing lines consisted of 1/2-inch and 7/8-inch hardline and were cut on a return loss bridge known to be accurate. The performance of the array had a familiar ring to it; Frank would receive excellent signal reports, but on receive, signals were far poorer than expected. Checking into the 432-MHz EME activity, Frank found that there had been a considerable number of other hams who had erected 16-Yagi EME arrays for 432 MHz that never worked well, and as a result, their stations had disappeared from the EME ranks.

At this point I began helping Frank to improve the array. The first priority, I decided, was to obtain a pattern measurement. The height of the array, 20 feet (6.1 meters) above ground, made the likelihood of taking accurate measurements remote. However, with the amount of array gain available (close to 26 dBd) I decided it should be possible to make adequate measurements by using Sun noise.[8,9] The E-plane pattern looked excellent, with the three major sidelobes down over 15, 25, and 17 dB, respectively



Six stacked NBS 17-element, 3.2 wavelength Yagi 144-MHz EME array.

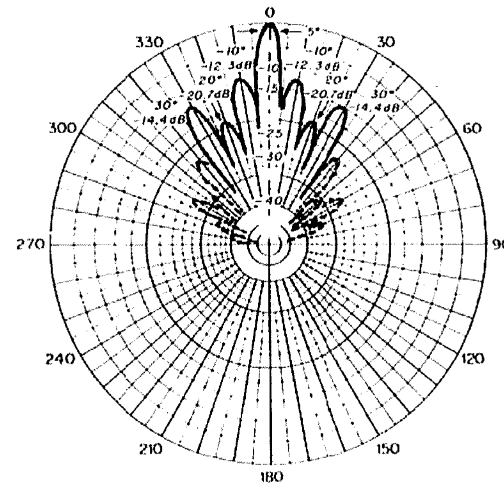


fig. 12. Sixteen stacked Cushcraft 424B Yagi H-plane pattern with 60 inch (2.2 wavelength) E-plane and 52 inch (1.9 wavelength) H-plane spacing.

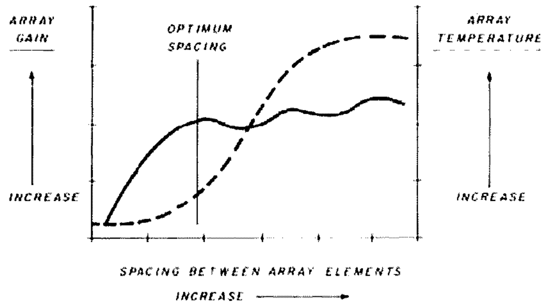


fig. 13. Array gain and array temperature versus spacing.

(**fig. 10**). The main lobe beamwidth was close to 6 degrees or greater than one quarter that of a single antenna. The H-plane was a shock with the −3 dB beamwidth of 4 degrees or much narrower than the expected 5.5 degrees. The major sidelobes were very large, at only 9.5, 11.5, and 14 dB below the main lobe (**fig. 11**). No EME signals had ever been copied when the array elevation was below 18 degrees. The fact that this angle was the same as the second sidelobe direction was no coincidence.

Some tests with a pair of K2RIW 19-element Yagis were run to measure changes in H-plane sidelobe levels. It was found that the first sidelobes changed at about 1 dB for every 2 inches of spacing change. Since the gain of the RIW19 (15.1 dBd) is close to the 424B it was felt that the results would be similar with the 424B. The vertical spacing was moved in by 6 inches (15 cm) to 52 inches (1.3 meters). The first sidelobes were expected to drop down to −12.5 dB. The results of that spacing change were amazing. The major sidelobes were now −12.3, −20.7, and −14.4 dB (**fig. 12**). Sun noise was up over 2 dB.* The on-the-air performance improvement was even more spectacular, with 41 QSOs made with 34 different stations during the first ten days of operation at the new spacing. Fewer than 20 QSOs were made in over two months of operation at the wider spacing. Most of the contacts were on random operation (as opposed to pre-arranged schedules) and included several 8 Yagi and one 4 Yagi stations. EME signals, including echoes, were now consistently copied down to 2 degrees elevation (the elevation angle where the main lobe would be clearing the earth). The main lobe beamwidth was still narrower than one quarter that of a single 424B at 5 degrees. This again supports previous measurements which indicated that a 1/4 beamwidth (5.5 degrees) would not occur until the first sidelobes are down 14 to 15 dB. The pattern indicates that the array has not yet been optimized. It is estimated that the best performance would occur at 62

*Sun noise is measured by pointing the array at cold sky, noting the noise level, and then pointing the array at the sun and measuring the noise increase. Sun noise is a combination measurement of overall receiver temperature and array gain.

inches (1.6 meters) E-plane by 50 inches (1.27 meters) H-plane spacing.

The reason for this dramatic improvement can be explained by looking at the approximate Earth noise pickup from the major sidelobes when the array was operated at low elevation angles. The main lobe at 432 MHz may typically see a sky temperature less than 20 degrees K. The sum of the first three sidelobes at −9.5, −11.5, and −14 dB would be equivalent to a single lobe 6.5 dB below the main lobe. Pointed at Earth (290 degrees K), those lobes would contribute about 65 degrees K of noise or cause a 5.1 dB degradation in signal-to-noise ratio. With the reduced sidelobes at the closer spacing, the sum of the three major sidelobes is about −9.8 dB and would cause about a 1.8 dB signal-to-noise degradation. This represents a *3.3 dB receive improvement*, which is quite significant on EME. The actual array noise is much more complicated; to calculate it would require summing all the sidelobes and accounting for the strength of the noise sources toward which they pointed. The over-2 dB Sun noise improvement nonetheless confirms the performance improvement.

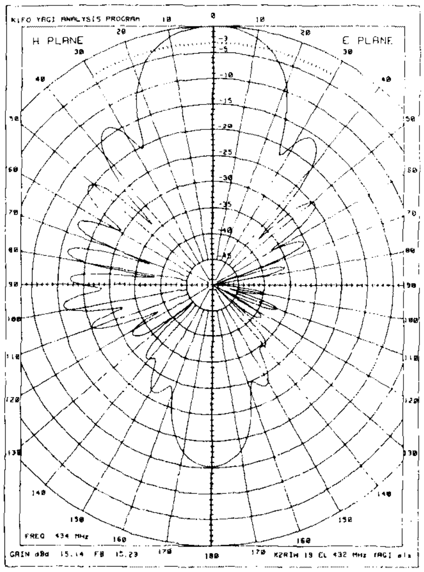


fig. 14. RIW 19-element Yagi computer generated E and H-plane patterns.

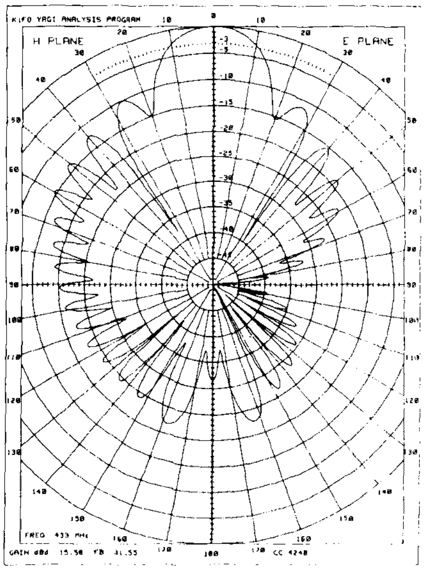


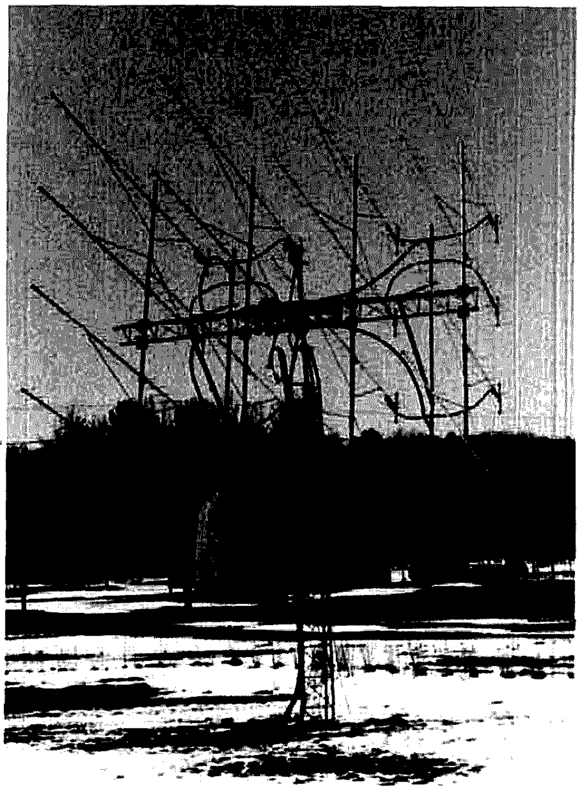
fig. 15. Cushcraft 424B Yagi computer generated E and H-plane patterns.

An alternative method for checking system temperature without taking array gain into account is to measure Earth noise. This is done by pointing the array at cold sky and then at the Earth and then comparing the noise levels. A well-performing 432-MHz EME system should see over a 4 dB ratio. Since this measurement does not take array gain into account it should not be used for array optimization because doing so could result in significant gain loss. **Figure 13** is a graph of typical array temperature versus stacking gain increase. The temperatures are not "real world" because they are quite different for the various Amateur frequencies.

easier method for measuring array beamwidths

Measuring the −3 dB beamwidths on high gain arrays with very sharp beamwidths can be a difficult task. WA1RWU's 16-Yagi 432-MHz array uses a prop-pitch motor for an azimuth rotator and has a 360-degree rotation time of over 3 minutes. The elevation leadscrew drive takes over 2 minutes to run 90 degrees. Even with such a rotating system, attempting to measure a 5 degree beamwidth can be a taxing experience. With a conventional commercial rotator it appears to be impossible to get an accurate measurement. It was discovered very early on that the half-power beamwidth was always equal to slightly less than 1/2 the spacing of the first nulls on either side of the main lobe. Interestingly, DL6WU has said that the beamwidth of a single Yagi is equal to 0.485 times the first null spacing. The measurement of the first nulls provides a much easier and most likely more accurate means of determining antenna half-power beamwidths.

The H-plane pattern measurements on the RIW19 Yagi provided yet another surprise. A pair of RIW19s had first H-plane sidelobes down 13 dB at a spacing of 54 inches (1.37 meters). The 424Bs had first H-plane



Sixteen stacked Cushcraft 424B Yagis in a 432-MHz EME array.

sidelobes down 12.3 dB at a spacing of 52 inches (1.32 meters). Considering that the 424B has an appreciably narrower H-plane beamwidth (22 inches versus 26 inches) and that it has more gain (15.8 dBd versus 15.1 dBd), just the opposite might be expected. The explanation again lies in the patterns of the individual antennas. **Figures 14** and **15** are computer-generated patterns for the RIW19 and the 424B. The RIW19 has, in general, sidelobes significantly lower than the 424B which would explain the different spacings for best array sidelobe levels. This effect has been noted in other Yagis also, and unfortunately makes most of the methods of calculating stacking distances (such as the aperture equations, or any of the various beamwidth formulas) of little use when array temperature is of concern. I spent a considerable amount of time trying to come up with a clever little formula for calculating stacking distances, but every one I tried had more Yagis disobey its numbers than ones that would.

This causes the optimum stacking distances to be different for different Yagis *even if they have the same gain*. Likewise a Yagi with a very clean pattern may have an optimum stacking distance greater than another with more gain but a poorer pattern. Keep in mind that "optimum" has, in this case, been defined as the highest G/T ratio.

mounting different frequency antennas on the same tower

The effect of having antennas for different bands located in close proximity has not received much attention. The interaction between antennas can be understood by looking again at **fig. 1**. If another antenna is located in the shadow of the first, it will not be able to extract as much energy as the first simply because the field strength is reduced behind the first antenna. To quantify the effect of having different antennas located nearby can be complicated. To evaluate the effect of a 144-MHz antenna on a nearby 432-MHz antenna, the aperture of the 144-MHz antenna at 432 MHz would have to be calculated in order to see how much of the capture areas overlapped. Next it would have to be determined who had "first dibs" on the signal or which antenna was in the other's shadow.

The only conclusive measurement I have been able to make has been to measure the Sun noise of a given array by itself and then add the other antennas and again measure Sun noise. I have found a consistent degradation in Sun noise by having arrays interlaced. Surprisingly, in most cases it is the lower frequency array that suffers. A side effect also appears to be the occurrence of stray sidelobes. The explanation of that phenomenon is that the unused antenna is re-radiating signals it had captured. Although not conclusive, terminating the unused antennas in a 50-ohm load appears to minimize the effect.

To be on the safe side, antennas should be located such that their apertures do not overlap. This can sometimes lead to very large spacings. As a practical matter the casual VHF/UHF operator may never see the performance degradation. The EME operator or enthusiastic weak signal worker who is looking for the last bit of performance is advised to either not mix arrays or to maintain sufficient spacing between them.

alternate stacking arrangements

This article has addressed only arrays with the Yagis arranged in uniform rows and columns. It may be possible to obtain additional stacking gain while controlling sidelobes by using other arrangements such as circle or diamond configurations. This would be due to the lower amount of aperture overlap required for sidelobe level control. I have not examined these alternative arrangements because of the difficulty in adapting them to an array with elevation control.

conclusions

Since most readers are likely to be more interested in a guide to how far apart to stack various antennas than in duplicating my work, **table 2** is provided. It covers a number of popular 144 MHz and 432 MHz antennas and their recommended stacking distances.

table 2. Measured performance of 144 and 432 MHz antennas.

144 MHz ANTENNAS

ANTENNA TYPE	GAIN dBd		PATTERN E x H	BOOMLENGTH	feet	SIDELOBES E x H -dB		STACKING feet E x H
6 el NBS	10.2	a	40 x 42	1.2	8.2	17	9	9.6 x 7.5
9 el F9FT	10.6		38 x 46	1.6	10.8	18	14	9.9 x 7.0
11 el Swan/KLM	10.8		40 x 44	1.8	12.3	13	10	9.6 x 7.5
11 el Cushcraft	10.8	b	40 x 46	1.7	11.8	19	13	10.0 x 7.5
12 el Swan/KLM	11.2		36 x 40	2.0	14.1	14	10	10.6 x 8.6
14 el Swan/KLM	11.8		34 x 37	2.5	17.4	15	10	11.0 x 9.5
20 el CC Colin	11.9		45 x 26	-.-	--.-	--	--	9.8 x 13.2
14 el Hy-Gain	11.9		35 x 35	2.3	15.5	--	--	11.0 x 9.9
12 el NBS	12.0	a	34 x 36	2.2	15.0	15	11	11.2 x 9.6
14 el Cushcraft	12.1		34 x 36	2.2	15.0	15	12	11.2 x 9.6
16 el KLM	12.2		29 x 31	3.0	20.7	12	9	11.4 x 9.6
11 el KLM 11X	12.2	u	34 x 38	2.3	15.3	19	14	11.5 x 9.9
16 el F9FT	12.5		32 x 34	3.0	20.8	17	13	12.2 x 10.2
11 el LUNAR	12.6	c,g	32 x 35	2.6	17.4	16	12	12.4 x 10.2
13 el KLM LBA	13.0		31 x 32	3.1	21.5	14	10	12.7 x 10.6
15 el Cue Dee	13.1		30 x 32	3.1	21.3	--	--	12.8 x 10.7
17 el NBS	13.1		28 x 33	3.2	22.0	13	10	13.0 x 10.4
19 el Cushcraft	13.2		28 x 33	3.2	22.0	14	11	13.0 x 10.5
13 el W2NLY	13.4	d	27 x 29	3.5	23.5	12	9	13.3 x 11.2
15 el Telrex	13.5		26 x 28	4.1	27.8	12	9	13.3 x 11.3
14 el K1FO	13.7	g	29 x 31	3.6	24.3	15	13	13.7 x 11.2
15 el NBS	13.9	a	26 x 29	4.2	28.7	14	11	13.7 x 11.4
16 el KLM LBX	14.3	u	28 x 30	4.1	28.1	17	14	13.8 x 11.6
18 el Cushcraft	14.5		27 X 28	4.2	28.7	15	12	14.0 x 11.8

432 MHz ANTENNAS

TYPE	dBd		E x H		feet	E x H -dB		inches
11 el Tilton	11.8	f,g	34 x 36	2.6	6.0	17	13	46 x 40
13 el W6QKI	13.3		27 x 29	3.4	7.8	12	9	53 x 44
13 el K2RIW	13.5		27 x 29	3.4	7.8	13	10	54 x 45
15 el NBS	13.9	a	26 x 29	4.2	9.6	14	11	55 x 46
16 el KLM LB	14.4		24 x 25	5.3	12.0	11	9	56 x 48
19 el K2RIW	15.1		24 x 26	5.6	12.8	16	14	60 x 54
21 el F9FT	15.2	h	24 x 26	6.6	15.0	14	12	58 x 52
26 el DL9KR	15.5	u,i	24 x 25	6.1	13.8	17	15	62 x 58
24 el Cushcraft	15.8		20 x 21	7.5	17.1	12	10	62 x 50
22 el DL6WU	15.8	u,l	23 x 24	6.9	15.6	15	14	64 x 58
24 el K1FO	16.6	k	22 x 23	7.5	17.2	16	15	66 x 60
28 el W1JR/DL6WU	17.0	u	20 x 21	9.3	21.1	14	13	70 x 64
30 el KLM LBX	17.3	u	19 x 20	9.6	21.9	15	13	72 x 66
31 el W1JR/DL6WU	17.5	u	19 x 20	10.4	23.7	14	13	74 x 68

a These NBS yagis have gain peaks 2 percent high in frequency.
b Gain peak is 11.1 dBd at 146 MHz, 38 x 44 degree pattern.
c Has incorrect balun length. With stock balun gain is 12.4 dBd.
d Figures are for 0.125 inch taper version with 20 in. reflector spacing
f Is tuned to 440 MHz. Retuned to 432 MHz gain would be 12.6 dBd.
g Design based on Greenblum / Tilton information.
h Is designed for 435 MHz. Gain peak is 15.5 dBd at 436 MHz.
i Uses 8 element screen reflector.
k Is a modified 424B using a single reflector and 22 directors.
l Is designed for 435 MHz. Gain peak is 16.0 dBd at 436 MHz.
u Design based on DL6WU information.

Although a few tenths of a dB additional gain may be obtainable at larger spacings, the added size, weight, and windload would most likely not justify the wider spacing even if EME or satellite communications are not anticipated.

It should be emphasized that gain alone does not

tell the whole story. A Yagi with a cleaner pattern may be a better choice for an EME array than one with higher gain and a "messy" pattern. The highest gain Yagi in the world is of little use if it splits into 5 pieces the first time a storm passes by.

Finally, the following summary should serve as a guideline in building multiple Yagi arrays:

- Optimum stacking distance is a compromise between gain increase and sidelobe level (G/T).
- Array −3 dB beamwidth will be equal to single element beamwidth divided by the number of elements in a plane when the first sidelobes are −14 to −15 dB. This usually represents optimum stacking or best G/T.
- The H-plane's inherently less directive pattern requires substantially closer spacing than the E-plane to achieve optimum sidelobe levels.
- Negating phasing line losses, doubling the number of elements in an array at optimum spacing will give approximately 2.7 to 2.8 dB gain increase in the E-plane and 2.5 to 2.6 dB in the H-plane.
- The greater the number of elements in an array, the more critical it is to have that array optimally spaced.
- The higher the frequency of operation, the more critical it is to have an array with small sidelobes (i.e. optimally stacked).
- The cleaner the pattern of an individual Yagi, the greater the optimum stacking distance will be — hence the greater the array gain.
- Although spacings closer than the maximum gain distance can cause the loss of a few tenths of a dB in array gain, the closer spacing can result in several dB of signal-to-noise ratio improvement on receive.
- If you are going to make a mistake, put your antennas "too close together" rather than "too far apart."
- Placing different band antennas in close proximity can degrade performance.

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ham radio

active antenna covers 0.5-30 MHz

Double duty allows 2-meter operation as well

Active antennas have been available commercially for several years, but relatively little practical information about them has appeared in the Amateur journals.[1,2] This article illustrates how to build a simple one covering from 500 kHz to at least 30 MHz using commonly available components. As an added bonus, the antenna works as an omnidirectional 2-meter antenna when not in use for HF reception. It will also meet the needs of SWL's for casual monitoring on the HF bands.

First, some clarification is in order. True active antennas (sometimes referred to as voltage probe antennas) use circuitry somewhat more sophisticated than what is presented here.[3] Antenna probe lengths of 2 inches have been used in some commercial models intended for military markets. Their circuitry does not match impedances in the usual sense, as does the simple approach used here.[3,4] Instead, their reactive components are made small in comparison to the extremely high source-follower FET input impedance, effectively swamping out the reactive component. A short probe antenna, at the lower HF frequencies, may have a capacitive reactance in the millions of ohms, while its effective radiation resistance is only a very small fraction of an ohm.[2] Despite its lack of eloquence the antenna presented here works well, and can be built in a weekend for very little cost.

general coverage antenna needed

Amateur Radio manufacturers have caught on to the market for general-coverage reception — today many transceivers will receive from the lower LF ranges continuously up to 30 MHz and beyond. While the benefits of general-coverage receivers are obvious, some problems await those seeking a single antenna capable of spanning several octaves.

The simple link-coupled LC preselector circuits used in vintage tube-type receivers were capable of matching to a wide variety of loads, and an "antenna" tuning control served in tweaking the most out of any antenna. But the situation is somewhat different today. Modern receivers are carefully designed for proper gain distribution. The tracking preselectors and high-gain front ends are gone; in their place are mod-

Internal view of unit shows the construction details of the active antenna. The ASP-677 commercial antenna was used. Because of the softness of the minibox aluminum, the 3/8-inch snap-mount washer was removed and installed on the inside wall of the minibox. This washer is part of the ASP snap-mount assembly.

By Peter J. Bertini, K1ZJH, 20 Patsun Road, Somers, Connecticut 06071

item	description
ANT-1	antenna probe, 20-50 inch length. Stainless steel antenna whip, aluminum tubing, or collapsible whip. Radio Shack No. 15-232 (38 inches) or No. 270-1401, 270-1402, 270-1403, or 270-1405.
F1	1 amp, fast-blow fuse
J1	RF connector, SO-239
RFC-1	20 turns No. 26 wire on Amidon FT50-61 core material (see text)
RL1	reed relay, SPST, 12 VDC coil.
S1	SPST switch, 125 VAC rating, Radio Shack 275-612 or equivalent
T1	9 bifilar turns, No. 26 wire on Amidon T68-8 core material. Wire 4-6 twists per inch (see text)
T2	power transformer, 117 VAC to 12.6 VAC at 100 mA minimum. Radio Shack No. 273-1385 or equivalent

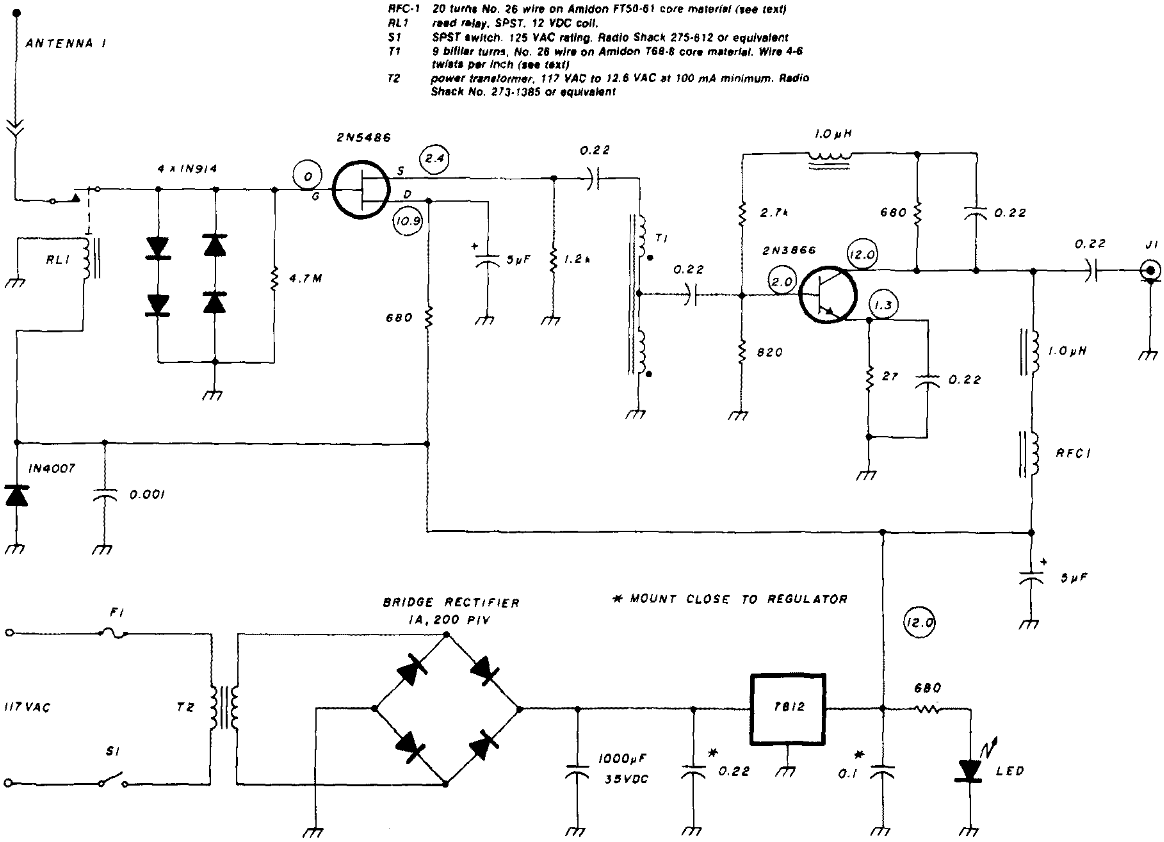


fig. 1. The schematic for the *indoor* antenna. This version has no provisions for 2-meter operation. DC voltage measurements are provided to assist in troubleshooting. *Voltages are indicated by circled values.*

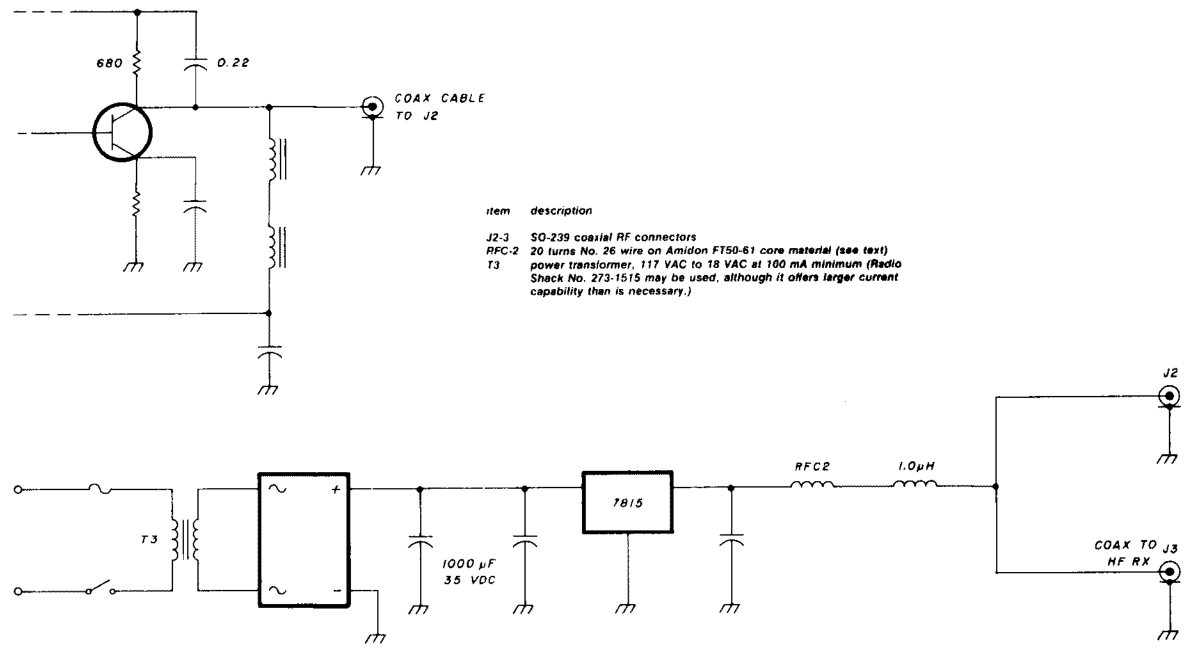
ern electronically-switched broadband filters. But these filters are not tolerant. If the load is not 50 ohms, their performance is no longer predictable and losses can become excessive. Consider the use of an 80-meter antenna for shortwave reception in the 40-meter region. An 80-meter dipole at 40 meters is virtually unusable because as the feedpoint impedance soars to a very high value the shunt capacity of the coax greatly attenuates incoming signals.

impedance transformation needed

Did you ever wonder how an automobile's AM antenna could work effectively on FM and yet be so short? The answer lies in the special high-impedance coax and special front-end design used in these receivers. (If you've ever tried to replace this coax with a length of RG-58 you've quickly learned a lesson in capacitive losses.)

If a short antenna will work at 550 kHz, it will work elsewhere as well. All that's needed is an impedance transformation network to change the extremely high termination impedance of our short "probe" antenna into a usable 50-ohm impedance for our receiver, across a range of several octaves. The size of an antenna has little bearing on its performance — full-size antennas do offer 50-ohm feedpoint impedances and low enough resistance losses to allow efficient RF radiation.[5] For receiving applications, many Amateurs have found that antennas such as the Beverage, ferrite-loop, and active antenna often out-perform full-size transmitting antenna arrays because of their lower susceptibility to noise pickup — and in some cases, their excellent directivity in the lower HF regions.

A JFET source-follower stage provides the desired high-to-low impedance transformation. (While a 2N5486 was used here, an MPF102 will also work.)



item	description
J2-3	SO-239 coaxial RF connectors
RFC-2	20 turns No. 26 wire on Amidon FT50-61 core material (see text)
T3	power transformer, 117 VAC to 18 VAC at 100 mA minimum (Radio Shack No. 273-1515 may be used, although it offers larger current capability than is necessary.)

fig. 2. The schematic for the outdoor antenna. Only circuit changes from *fig. 1* are shown. Note the elimination of the 0.22 μF coupling capacitor, allowing DC feed on the coaxial cable. Power supply components have been changed for 15-volt output, compensating for possible resistive cable losses. Component values not shown are the same as in *fig. 1*.

The antenna is a length of wire, about 48 inches (1.22 meters) long. A length of stainless-steel wire salvaged from an old VHF whip or automobile antenna or even a section of aluminum tubing may be used for the antenna probe.

Two versions of this antenna were built and tested; one was designed for indoor use and the other for outdoor mounting. The indoor unit was built in a small enclosure. The exact dimensions are not critical, providing all components can be comfortably mounted inside. The only difference between the indoor and outdoor models is in the packaging — the outdoor model uses a remote weather-proof housing for the preamp and antenna probe. Everything is self-contained in the indoor version.

Referring to **fig. 1**, for the schematic for the indoor model, note the back-to-back diode protection from the JFET gate to ground. I found that the static discharge generated from shuffling across the shack floor was responsible for the untimely demise of several 2N5486 devices before I realized what was going on. These diodes do not guarantee total protection — the best safeguard is to avoid touching the antenna probe. To prevent generation of an insidious form of TVI in the outdoor antenna caused by diode rectification of RF radiated from nearby HF or VHF transmitters, a small reed relay is used to disconnect the antenna probe from the preamp stages when the antenna is not in use. This also offers some protection to the JFET during electrical storms, when static charges might build up on the antenna. The reed relay should be used in the indoor model as well for protection against static discharges. The reed relay is used only in the HF model — for 2-meter operation a more involved switching scheme using DPDT relays is needed. (This will be addressed in greater detail below).

The schematic shown in **fig. 2** serves as the basis for the outdoor HF active antenna. The only difference is in the power feed arrangement. In the outdoor model, power for the preamp head is fed through the coax from the indoor power supply.

additional gain needed

Following the JFET device, which does not provide voltage gain, is a class-A bipolar stage using a 2N3866 transistor. (A 2N5109 could be used here instead.) "Hotter" transistors such as the TRW LT1001 might offer some improvement, especially at higher frequencies, but they might also require some circuit revisions to prevent self-oscillation. This stage does provide gain, building up the very small signals captured by the short antenna. The transistor must be heatsunk; the stage draws about 25 mA resulting in considerable

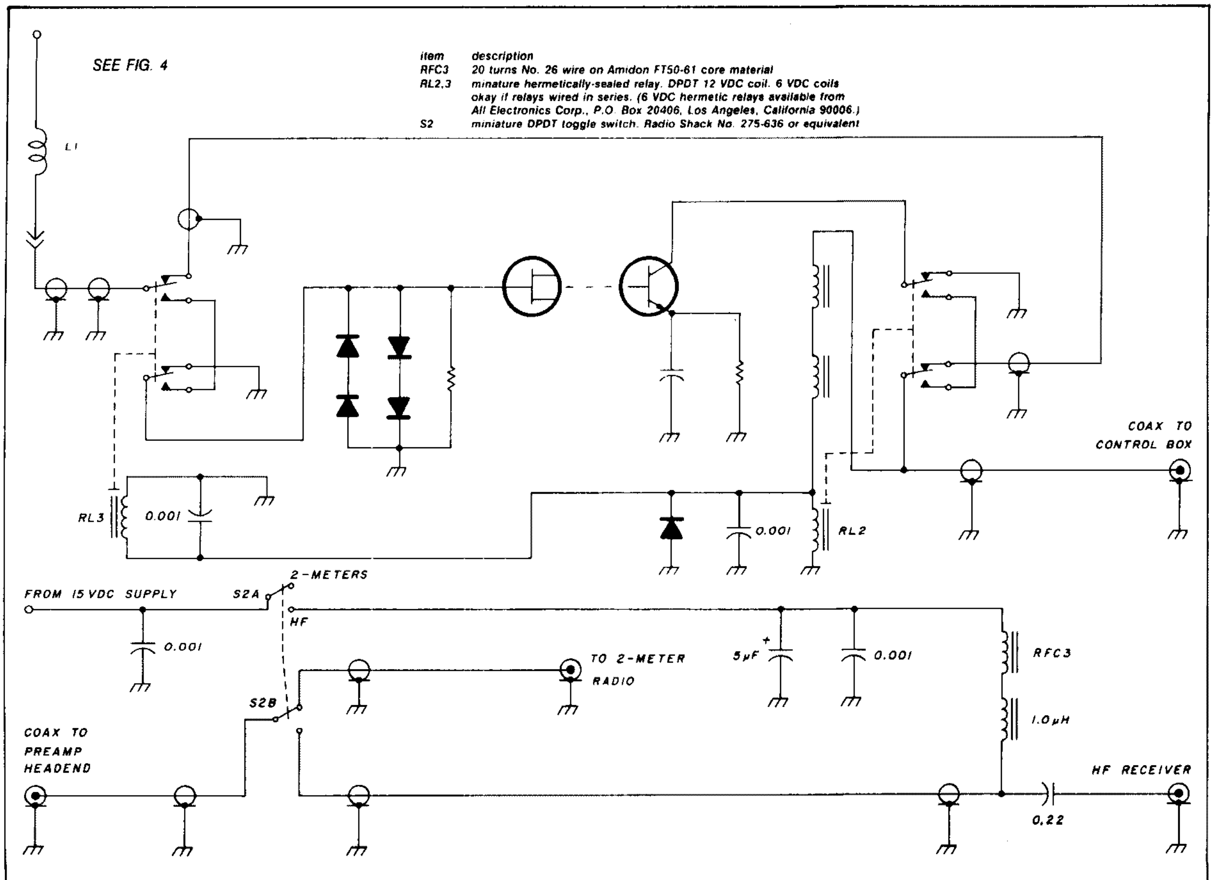


fig. 3. The changes allowing 2-meter operation for the active antenna while not in use on HF. The 15-VDC power source is the one shown in *fig. 2*. Two DPDT relays bypass the active antenna preamp stages for straight-through operation on 2 meters. Six-volt relays will work, providing the coils are wired in series. Component values are not shown if unchanged from previous diagrams.

device dissipation. Shunt feedback sets the stage gain and insures stable operation. A series choke in the feedback network allows greater gain at higher frequencies, this is necessary because the f_T (the gain-bandwidth product, or the frequency at which the current gain reduces to unity) of the 2N3866 is only 500 MHz. A broadband interstage matching transformer is used between the source follower and bipolar amplifier. Using 20 bifilar turns allows enough inductive reactance for operation through the AM broadcast band. If AM broadcast band coverage is not desired, the windings may be reduced to nine bifilar turns. This will cause gain to start rolling off below 1800 kHz, while yielding a slight improvement above 30 MHz.

All models were constructed and tested using point-to-point wiring techniques on double-sided PC board material. Phenolic soldering strips provide the mechanical support needed for components and wiring. While the results are not overly photogenic, the circuit works as intended. Etched printed-circuit construction is not needed here.

The outdoor model works best.[1] Mounted high and in the clear, it is not subjected to the man-made noises carried and radiated on the home's wiring. Many common home appliances will generate copious amounts of broadband hash across the HF spectrum: vacuum cleaners, television horizontal-sweep circuits, fluorescent lights, light dimmers, computers, and numerous other devices will cause grief for the serious HF operator. While the antenna can be side-mounted to an existing tower leg, it is best to mount it away from nearby metal objects. If at all possible the antenna should be top-mounted on the tower masting. Lead-in coax can be either RG-59 or RG-58 type cable. RG-8 coaxial cable can be used with the outdoor combination 2-meter/HF model to limit the VHF losses.

use for 2 meters as well

Since the antenna probe is 48 inches long, we have the basic foundation for a 5/8-wavelength vertically polarized antenna for 2 meters. All that's required is a matching transformer (**fig. 4**) to match the

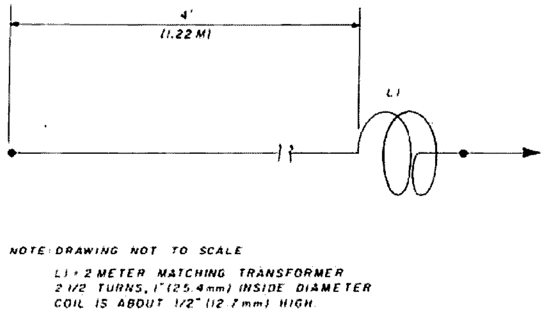


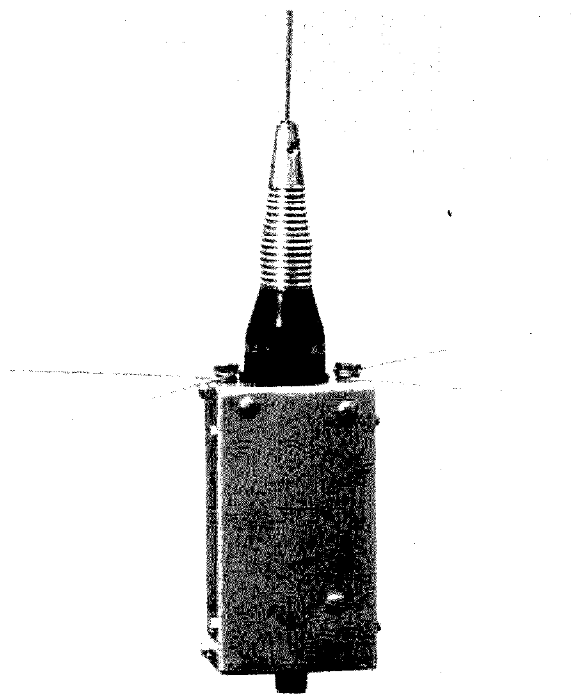
fig. 4. Construction details for the 5/8λ two-meter radiator and matching transformer. The transformer may be formed by winding the antenna rod around a suitable form. These dimensions were taken from the ASP-677 matching transformer assembly.

5/8-wavelength radiator to 50 ohms. Given a length of stainless steel antenna rod, it is possible to make the coil and whip one continuous piece. A commercial 5/8-wavelength antenna, such as the Antenna Specialists model ASP-677, could serve here as well, and was my choice for my outdoor HF/VHF combination antenna. Note that the ASP whip is cut for about 46 inches (1.17 meters); this is because of the added length introduced by the spring assembly. *Half-wave or 5/8-wavelength antenna designs using grounded, tapped matching transformers will not work.* The extremely small inductances will short circuit HF signals.

Three or four ground radials are needed for the 2-meter/HF combination active antenna. These should be a little longer than a quarter-wavelength, 20-inch lengths of stainless steel antenna whip wire or aluminum tubing will serve here. SWR is adjusted by trimming the radiator length; 48 inches will be about optimum for the upper 2 MHz of the band. If trimming the antenna does not yield an SWR under 1.5 to 1, compressing or expanding the transformer coil should provide a good match. The radials must make good electrical contact to the metal enclosure housing the antenna. A Radio Shack catalog No. 270-238 mini-box houses my unit. Its 5-1/4 × 3 × 2-1/8 inch (13.34 × 7.62 × 5.52 cm) size is roomy enough to comfortably mount the preamp components. The PC board used for mounting the active antenna components must also make a good low-impedance RF ground connection to the enclosure. This provides the needed RF ground path for the 2-meter antenna. If the ASP antenna is used, a short length of RG-58 coax should be used between the 3/8-inch snap-mount and preamp changeover relay. Use short coax lead connections to the relay and ground.

Figure 3 outlines the electrical details for the dual-purpose antenna. Two miniature hermetically sealed relays are used to bypass the preamp stages, allowing straight-through operation for 2 meters when the power is removed from the antenna. Metal-can 6 VDC hermetic relays can be used, providing the coils are connected in series. Six-volt relays are often more easily available, and cheaper, than their 12-volt counterparts. A 1N4007 silicon power diode protects the amplifier from reversed supply voltages; and from the counter-EMF generated by the relay coils as power is removed. Depending on the length of coax used, the power supply voltage may have to be increased to compensate for voltage drops in the coax cable. The 2-meter version draws over 100 mA; at least 12 volts should be available at the antenna preamp.

Short leads are mandatory for good VHF operation. The straight-through 2-meter operation uses a short piece of coax cable between the relays. RG-174 is okay for this, but extremely short ground and center wire connections must be made at the relays. The shack-mounted power supply unit should be as carefully constructed, to keep the losses at a minimum. Note the use of the DPDT switch; it does "double-duty" in switching the power to the antenna preamp and also changing the feedline between the HF receiver and



The finished dual-purpose active antenna. Note the installation of the 1/4-wave 2-meter radials. (See W6SAI's November, 1984, column in *ham radio* for a timely discussion on improving radial performance for 5/8-wave 2-meter antennas used in this fashion.)

2-meter FM rig. Very short leads are used for the coax connections at the switch terminals, avoiding severe impedance bumps at 146 MHz. Use care in soldering; overheating the center insulation may result in a short to the coax braid. The outdoor antenna enclosure should be weatherproofed. (Silicone bathroom caulk will do well for this purpose.) Be sure to allow for condensation; a small drain hole on the bottom of the enclosure is needed. Also spray the PC board and components with clear acrylic spray to prevent corrosion. Anti-moisture and fungicidal varnish, carried by some electronic suppliers, is what I used, and it is also ideal for sealing the coax connections after they have been taped.

what to expect

Although performance of either model is adequate for casual monitoring, it is difficult to evaluate these antennas unless an antenna test range equipped with full-size antennas for all of the frequencies involved is available. S-meter comparisons alone can be deceptive. A higher S-meter reading may be obtained with an active antenna with excessive internal gain, while the signal-to-noise ratio may be best on the comparison antenna. This lesson was demonstrated some years ago when I tried using an old RME preselector ahead of a 51J3 Collins receiver. Above 15 MHz the S-meter was very "busy". Switching off the preselector often dropped S-9 signals to an S-1 or less, even though the readability increased dramatically.

With the active antenna connected to your HF receiver and turned on, a slight increase in receiver noise should be heard. Disconnecting the whip antenna from the preamp should cause a noticeable reduction in receiver noise. This is a good indication that everything is working as it should. The gain of this active antenna is not as great as some commercial units I've used in the past. Do not expect signals to produce the same S-meter readings as a full-size antenna.

Some variations of the design are possible. SWL's with a need for coverage in the LF ranges below 550 kHz might consider eliminating the interstage matching transformer. However, without the matching transformer, 60-cycle cross-modulation from strong nearby AC powerline fields will be an annoying problem. This can be especially troublesome in the indoor model where the probe and preamplifier stages are near to the power transformer. The limiting factor for LF performance is the choke values used to isolate the DC supply voltages. They will have to be increased to allow operation below 500 kHz. There is no magic in the antenna probe length — antennas as short as 20 inches might prove adequate. The 48-inch length was selected to allow 5/8-wave 2-meter operation. An 18-inch (45.7 mm) probe could serve as a 1/4-wave 2-meter radiator and eliminate the need for the 2-meter matching transformer.

A few words of caution: the 2-meter antenna will handle about 25 watts. More than this invites damage to the 1-μH chokes. When using the antenna on an HF transceiver, never transmit into the antenna. Always disconnect the mike and key and set the drive and mike controls at minimum *before* connecting to the active antenna!

There you have it, all in one package: an all-band HF antenna for SWL'ing on your transceiver and a 2-meter antenna for your FM rig. Considering the obstacles facing many Amateurs — lot size, restrictive zoning regulations, and unsympathetic landlords — this antenna will offer a lot of performance in a small, inexpensive and inconspicuous design.

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ham radio

SEE YOU AT DAYTON

control your take-off angle: the JR vari-lobe antenna

Dial in your skip distance and maximize your signals

The *ham radio* staff was saddened to learn of the recent death of Dick Schellenbach, W1JF.
In memory of this exemplary Radio Amateur we are proud to present the following two articles.

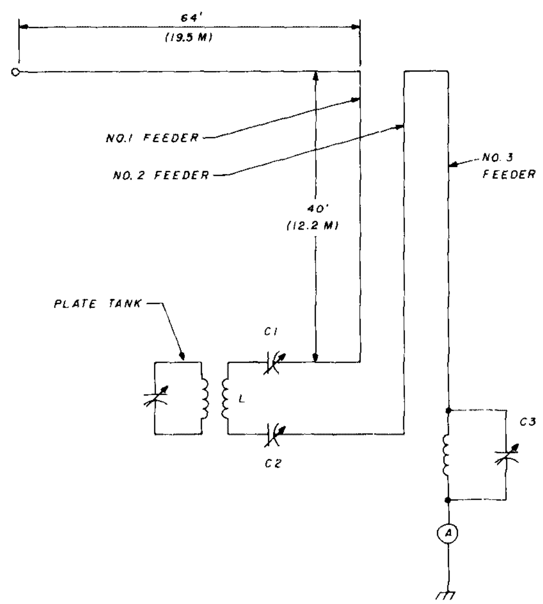

fig. 1. *The original Reinartz antenna system. A current-fed Hertz (on 80 meters) with one quarter wave section horizontal and unconventional three-wire tuned feeders.*

Most Radio Amateurs, I'm sure, are quite familiar with the directional properties of simple halfwave antennas. In these antennas, the majority of radiation is broadside to the radiating element, in the case of a horizontal type antenna. Some Amateurs may be familiar with, or may have experienced, the effect of raising or lowering an antenna and have seen how its launch angle, as well as its impedance, was significantly affected by the change in height. In this respect, a lower angle of vertical radiation usually produces a stronger signal further out in distance than one that has an extremely high angle of radiation when the communication path is governed by sky waves. This path distance versus launch angle phenomenon is very noticeable on the lower frequency Amateur bands of 160, 80, and 40 meters, where both high and low angles of radiation are useful in radio communications.

Given the fact that a half-wave antenna generally radiates most of its signal broadside to the antenna, what would you think of a simple modification that would permit controlled selection of its radiation angle? This would allow the Amateur station to take advantage of the prevailing ionospheric conditions, night or day, and therefore enhance signals to and from a particular distant location. Such an antenna system *is* feasible.

The basic idea for the variable lobe system was developed fifty years ago by the legendary John L. Reinartz, W1QP.[1]

In developing this new concept and publishing the results of his experiments Reinartz discovered he truly had a multiband antenna system as well as one providing versatile radiation angle control allowing him to maximize signal strengths at different locations. Reinartz also found out he could compensate for time-varying propagation conditions, especially when the

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reflective layer was in the process of changing heights, thereby requiring a different radiation angle while he was communicating.

The ability to alter an antenna's vertical lobe characteristics is one solution to the thought-provoking question, why does it often seem that signal reports appear to differ drastically between individual stations, even when the stations are located only a few miles apart in the same general direction?

The causes for inconsistencies between signal reports include a myriad of possibilities such as polarization differences, vagaries of the ionospheric reflective layers, multi-hop propagation, launch radiation, and receiving station antenna angle differences, to name a few. Because of the large number of variables affecting sky-wave propagation, the solutions to these problems are complex. However, if a radio station possesses the capability to adjust to any potential condition, the inconsistencies are not at all important — only the corrective measure capability matters.

Reinartz recognized the same perplexing propagation differences between the station he worked and proposed that if he were to vary the vertical angle pattern of his antenna, he should maintain a maximum signal strength at both ends of the communication path. In his model, Reinartz developed an antenna system which is basically a simple current-fed Hertz type antenna with one quarter-wave horizontal section and another sixth-wave feeder section, the latter consisting of two normal operating feeders and a third, vertical wire carried into his ham shack for tuning and compensation control. The original Reinartz version of the controlled radiation angle antenna is shown in **fig. 1**. The model as shown, operating on the 80-meter band, proved to be capable of operating under vertical angle of radiation control at twice or even four times its fundamental design frequency.

In spite of the antenna's multiband capability, it was used primarily on the 80-meter band, where daily and seasonal propagation effects were more pronounced and allowed Reinartz to do a lot of experimentation during variable band conditions.

how it works

Referring again to **fig. 1**, illustrating the original Reinartz antenna system, it should be mentioned that the variable capacitor C2 was found to be unnecessary in the 80-meter application and that all vertical angle lobe changing was affected by varying C1 while maintaining resonance of the parallel L/C network against ground for each particular operating frequency. Reinartz used an RF current reading ammeter to determine resonance in the third (vertical) feeder wire, but the meter is deemed unnecessary, as other means are available to determine proper resonance of the vertical wire section.

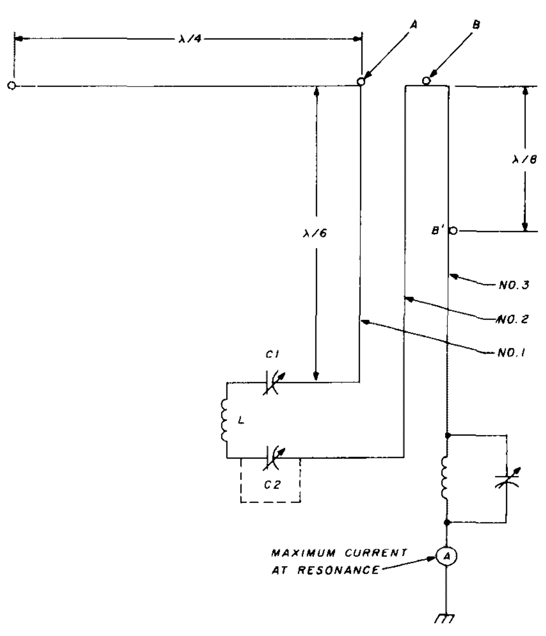

fig. 2A. Mechanism for changing current point *B* up or down by adjustment of series capacitor C1.

More details of the mechanism for changing the antenna's vertical field pattern is shown in **figs. 2A** and **2B**, where on the 80-meter band these figures indicate the voltage and current distributions on the model and the method employed to change the radiating current points on the system. The horizontal wire section of the antenna in **fig. 2A** is only one-sixth wavelength above the ground and hence, most of its radiation is directed upward — a real "cloud burner." During nighttime conditions, for example, this pattern would produce strong local signals because of the high angle and short distance skip zone; if we were to move the current point from the top of feeder number 3 downward, however, then the radiation characteristic would be reversed and the antenna's radiation would become predominately low angle — good for DX.

The change in vertical radiation pattern with relationship to the RF current maximum point on the third (vertical) wire section is illustrated in **fig. 3**. As one may surmise, the third vertical wire section of the antenna functions to provide the vertically polarized and low angle lobe control for the system. By controlling the current distribution by altering capacitor C1 in the series tuning circuit of **fig. 2**, the Vari-Lobe antenna changes from a high vertical angle radiator to a low vertical angle radiator quite simply. When capacitor C1 is slowly tuned, the antenna's vertical pattern may be finely adjusted anywhere between the two extremes dipicted in **fig. 3**. This tuning control

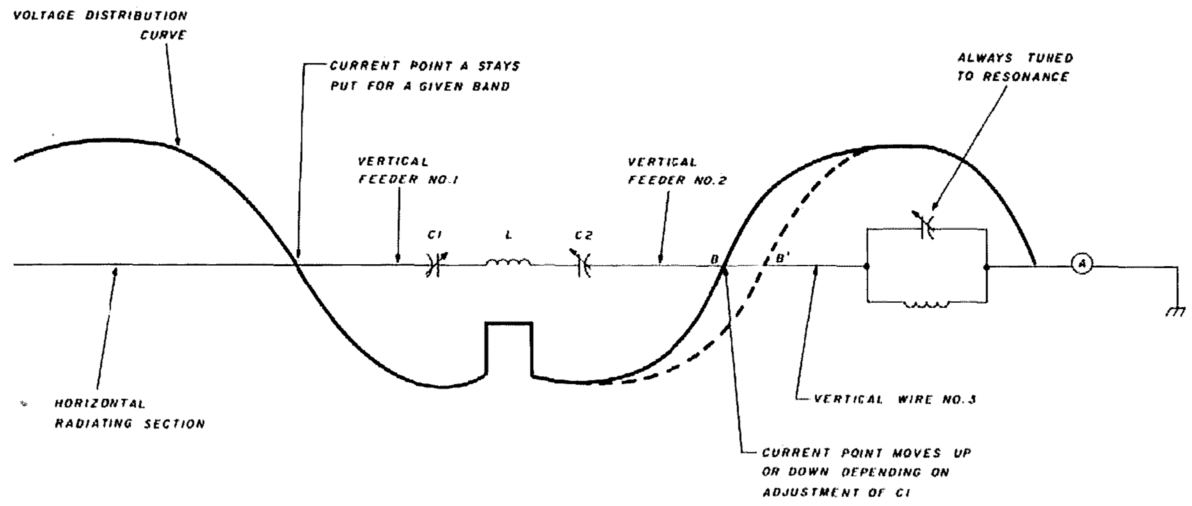

fig. 2B. The antenna system shown stretched out to illustrate the standing wave (voltage and current) distribution (80-meter example).

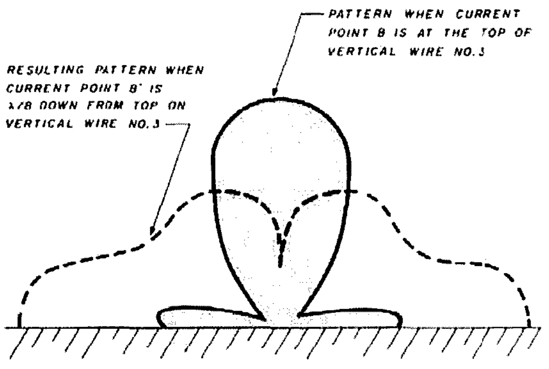

fig. 3. The vertical field pattern differences with shifting current point *B* to *B'*.

effectively allows the operator to "tune in" the required skip distance to another station for maximum signal strength. Large departures in current control by changes in capacitor C1 will require slight readjustment to the resonance of the parallel tuned circuit between the vertical section wire number 3 and ground.

According to tests by Reinartz in which he instrumented the vertical wire from top to bottom with small pilot lamps to indicate antenna current distributions, the further downward the current maximum was toward the one-eighth wavelength point, the greater the low angle effect and the further out the skip became.

It is interesting to note that Reinartz performed many of his early experiments and collected data on the shifting field strength patterns on the 10-meter band. Later, actual on-the-air tests using a full-scale 3.5 MHz (80 meter) model was used to communicate with a number of Amateurs up and down the east coast of the United States.

Concluding remarks by Reinartz indicate that his repeated tests showed that compensation for the advance of time from early evening to past midnight must be made by a change in the location of the maximum current point on the vertical section in order to maintain maximum signal strength at the receiving station. During daytime operation, he noticed that although there was one generally satisfactory setting of C1, slight changes could be made in the lobe adjustment to provide a maximum signal at some particular receiving station. He noticed too, that some stations were always consistent in the antenna's lobe setting for obtaining a maximum signal strength.

a modern version of the JR antenna

By carrying Reinartz's ideas further and adding modern improvements, the JR Vari-Lobe antenna came into being. The new system offers more operating frequencies with ease of controlled radiation from either the ham shack or by remote control. The modern version is shown in **fig. 4** and its control section in **fig. 5**. I prefer the remote tuning method because it keeps the necessary radiating elements away from the surrounding objects such as houses, trees, telephone and power lines, etc. The JR Vari-Lobe antenna's performance could be adversely influenced by such local obstructions because the antenna is a relatively low height system and it utilizes low angle vertical polarization.

As an additional feature of modernization, I incorporated a 15-meter "high gain" capability by using the "JF Array"[2] principally on the horizontal section. When used, the JF Array mode is automatically invoked when the system is tuned up on the 15-meter band. On this frequency, the antenna acts as a two-element, wide-spaced collinear array producing a good 3-dB gain broadside to the horizontal sections. Vertical angle selectivity is not used on this higher frequency band because the system is operated in its "Zepp" mode, in which feeders A and B are used and B and C are tied together as shown in figs. **4** and **5**.

The reader will notice that there are small differences between the original Reinartz antenna and the modern version. Most of the recent changes were brought about by the addition of the JF Array feature and a more compatible feeder length to handle multiband applications. In a departure from the original antenna system, we have taken advantage of using a multiband tuning network to implement the third wire to ground resonance of the system. This expedient is in keeping with modernization of the compensation tuning, since in practice, only one adjustment needs be made for each operating band segment, and one that holds for a bandwidth of 40 to 60 kHz on the 80-meter band and greater than 60 kHz on the 40-meter band.

The coaxial transmission line to the radio station equipment may be provided by either a 50 or 75-ohm coaxial cable through a 1:1 balun transformer connected across a few turns at mid-point on inductor L of **fig. 5**. As discussed previously, capacitor C1 of **fig. 5** is the vertical lobe angle control, where decreasing its effective capacity corresponds to raising the current maximum point on the vertical wire number 3 corresponding to a high lobe angle for the antenna. Conversely, an increase in capacitance by C1 results in lowering the current maximum point on the vertical wire number 3 corresponding to a low angle of radiation for the antenna. Each extreme capacitance change of C1 must be followed by an adjustment of the parallel tuned resonance circuit, especially if the operating frequency has also been changed beyond the bandwidth restrictions previously mentioned.

performance

Because the JR Vari-Lobe antenna system is intended to be used primarily on the 80 and 40-meter bands with infrequent excursions to 15 meters, it is important to have a good ground in order for the system to work well. Ground radials should be employed beneath the horizontal section of the antenna and extend outward in a fan shape from the third (vertical) wire section. A good ground system will not only enhance the lobe selectivity feature, but will also significantly diminish ground losses surrounding the radiating system, especially on the lower frequency bands.

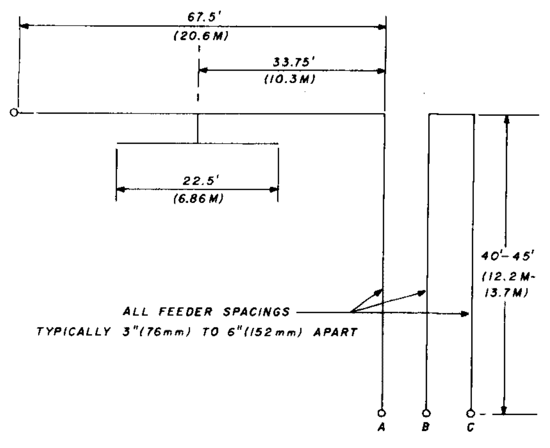

fig. 4. The modernized JR vari-lobe antenna system.

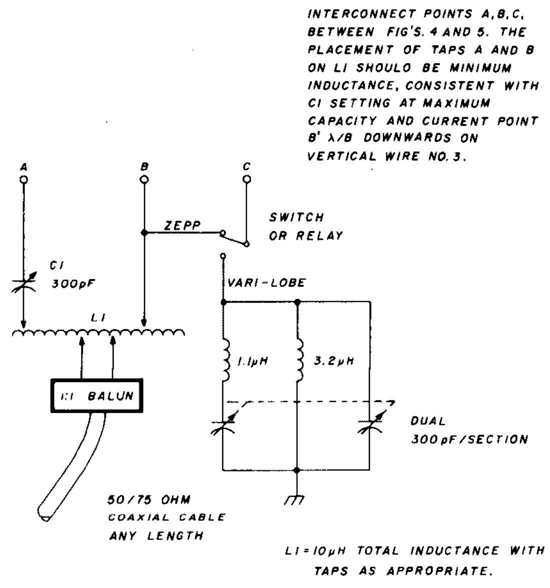

fig. 5. The tuning section for all bands and modes.

On-the-air tests were conducted on the 80, 40, and 15-meter bands over a period of three months. Based on these tests, it may be stated that the JR Vari-Lobe radiation characteristics on the 80 and 40-meter bands successfully proved to correspond to Reinartz's theory. The variable angle selectivity feature has enabled maximum signal strength adjustment over different path distances both during night and daytime conditions. While operation on the 15-meter band does not include variable angle control, it has proven to be an effec-

tive radiating system of modest size and relative simplicity. Although tests were not performed on the 10 and 20-meter bands, Reinartz has shown the antenna system will operate as a conventional Zepp on all bands and if desired, with variable angle control as well. Conventional Zepp configurations do not require the parallel tuning network to ground for the third (vertical) wire. Thus, the simple expedient of tying together feeders B and C of **fig. 5** changes the antenna into a Zepp configuration. The Zepp mode may be used on all bands higher than 80 meters, providing a flexible multiband antenna system if you don't mind the coupling adjustments and tuning of two feeder wires. The system is a natural for the Amateur who has limited antenna space but would like something different and more versatile than other systems.

resonance tuning features

The original Reinartz antenna system used an RF ammeter in series with the ground connection to determine resonance of the antenna when excursions were made in lobe angle and frequency. A diode detector probe positioned near the top of the parallel tuning network will work as well and could be remoted quite easily if desired. However, it is also posssible to tune the system "by ear." When the parallel tuned circuit is tuned to the correct resonant point, a noticeable increase in received signal strength or background noise is apparent, thereby dispensing with the requirement for remote meter reading.

The multiband tuner operates satisfactorily over all of the harmonically related Amateur Radio bands to maintain resonance of the system. Obviously, if only limited operation is contemplated on 80 or 40 meters, then the multiband tuning function may be eliminated and a single-band network would serve as well. In my installation, using remote controlled tuning proved to make the operation of the Vari-Lobe a joy to use. Small 24-VAC furnace valve control motors were coupled to the shafts of both tuning capacitors to provide remote tuning control from the shack. These small, economical motors rotate slow enough to allow sufficient tuning resolution and accuracy even though they are turning only in one direction. An alternative to these AC-type control motors are surplus stepping motors featuring bi-directional incremental DC pulse control with 3.65 degree/step resolution. These devices are available from a number of sources for a nominal price. Some even include additional gearing for slower response.

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ham radio

the end-fed 8JK: a switchable vertical array

30/40-meter antenna radiates at low angles and requires little space

This array was developed to provide maximum gain in a compact, convenient low-vertical angle antenna system. The 40-meter version requires only 1225 square feet of area, and the 30-meter version, less than half that amount.

The 30- and 40-meter bands were selected because of the increased activity on those bands — especially for DX — resulting from the diminishing MUF caused by declining solar activity. The selectable directivity and low radiation angle of this system should offer significant performance advantages as interference and competition on these bands increase. While the gain of this phased array is modest, the system will effectively double the RF radiated power of a single vertical in the preferred direction. The array also provides deep nulls spaced 90 degrees apart on each side of the main beam pattern, thus improving rejection of unwanted signals.

theory of operation

The 2 x 2 phased array is a pair of two parallel-element end-fire vertically polarized matched arrays, each of which is similar to W8JK's single section beam, with elements fed 180 degrees out of phase and spaced 1/8 wavelength apart. A composite simplified view of the array is shown in **fig. 1**. By folding the top and bottom 1/16-wavelength end sections of each element, a useful height reduction results without significantly effecting similar normal full-size array characteristics caused by element folding. Because each connecting wire at the base and the elevated end sections are only 1/16 wavelength long and carry very little current, little radiation and pattern distortion result from these folded sections

construction and implementation

Figure 2 depicts the 2 x 2 phased array configuration; **table 1** provides the element lengths and spacing as well as mast height requirements. **Figure 3** illustrates the matching 1/4-wavelength stub and

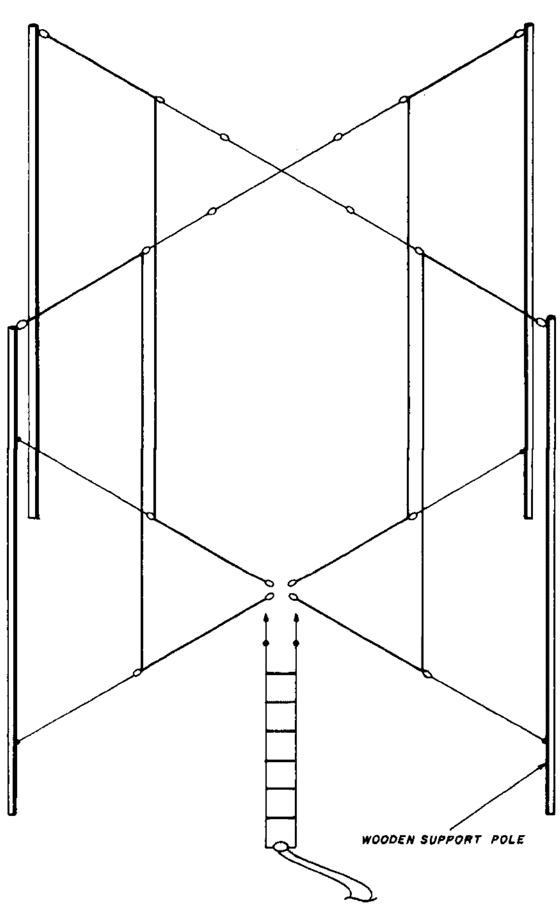

fig. 1. Composite — simplified view of 2 × 2 phased array.

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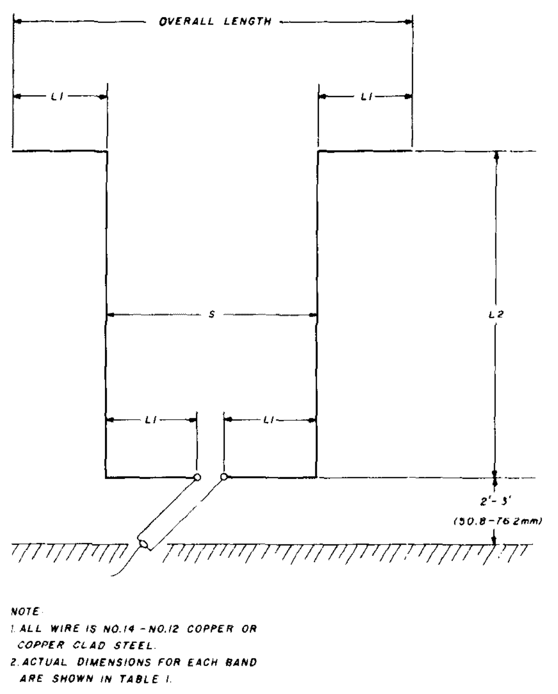

fig. 2. 2 × 2 phased array configuration; single element shown.

remote relay switching scheme including a balun for mating with coaxial cable. The length and physical positioning of the coaxial cable into the station are not critical.

Because the 2 x 2 phased array consists of two identical two-element arrays crossing at right angles to each other, a simple DPDT relay controls the selection of the array to be connected to the common 1/4-wavelength stub and hence through the balun transformer and coaxial cable into the station.

Once the 1:1 balun transformer tap position has been adjusted for a low VSWR at the desired operating frequency, no tuning or matching are required between the station and the system. The actual tap position depends on the stub line impedance and interwiring characteristics of the relay switching.

In order to reduce difficulties in array matching as much as possible, each element on an array — as well as the overall symmetry of the arrays — should be made as identical as possible to preserve balance and uniformity when changing directivity. Because a common 1/4-wavelength stub is used for matching into the antenna system, it is important that all wire connections to the DPDT relay be made equal in length. In placing the tap point for the balun transformer, you will find that with 75-ohm coaxial cable, the tap location will be a few inches higher toward the relay end than that of 50-ohm coaxial cable. More detail on proper matching is provided later, but the process is simple and provides wide bandwidth on either band of operation.

table 1. Physical dimensions for the 2 × 2 phased array on 40 or 30 meters.

	40-meter operation	30-meter operation
L1	8.75 feet	5.80 feet
L2	43.00 feet	29.00 feet
L3 (stub)	34.00 feet*	23.25 feet*
S	17.50 feet	11.60 feet
overall length	35.00 feet	23.25 feet
pole height (min.)	45.00 feet	31.00 feet

*May be laid horizontal as required.

site and space requirements

Four poles, preferably wooden, are required to sup-

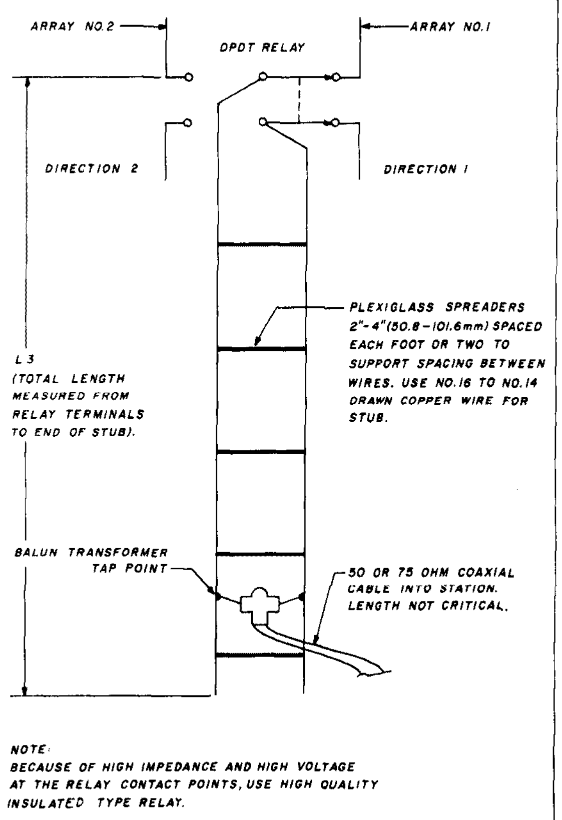

fig. 3. Matching quarter-wave stub, relay and balun transformer installation.

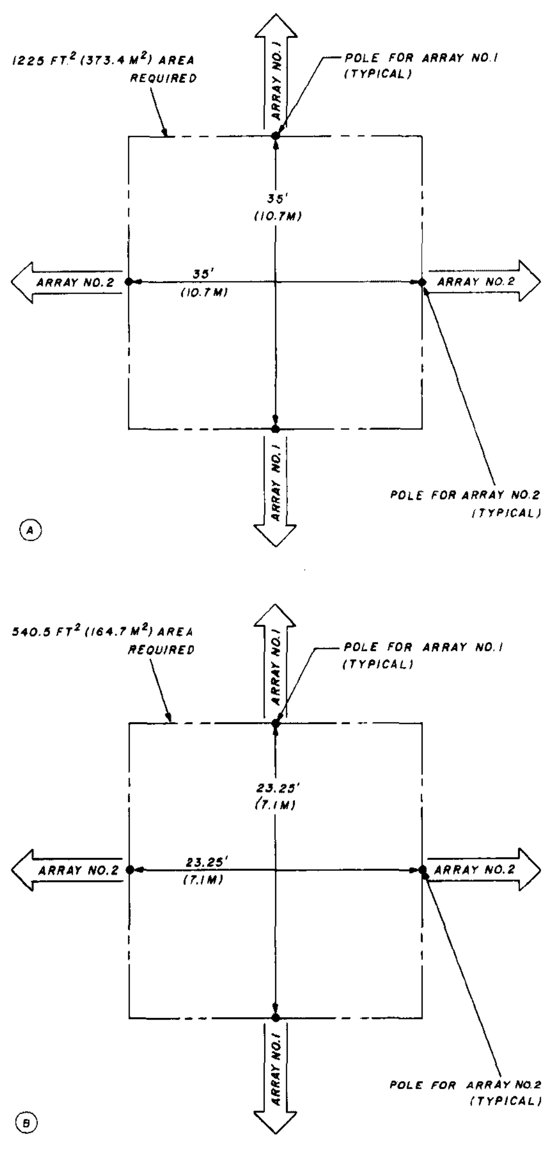

fig. 4. Site and space requirements of 2 × 2 array: (A) 40-meter version and (B) 30-meter version.

port the array. While they are oriented as shown in **fig. 4**, the distance between them can be increased to meet individual requirements. **Figure 4** also indicates the principal beam directions with respect to the DPDT relay position that is controlled from the station end. Using this directivity characteristic of the array, the individual beam patterns may be oriented in the most favorable compass directions.

To assure maximum low angle radiation and deep null beam pointing performance — and to eliminate effects of pattern distortion and RF absorption — the array should be located in an area free of structures or power lines for at least 250 feet from its center on 40 meters and 190 feet on 30 meters.

technical performance

The 2 x 2 phased array provides approximately 4 dB gain over conventional 1/2-wavelength dipoles and 6 dB gain over 1/4-wavelength ground mounted single verticals. The azimuthal bi-directional beam pattern, although relatively broad, (approximately 80-degree half-power beamwidth) has quite deep nulls, better than 20 dB down from the main lobe maximum. The nulls are also bi-directional in either case and are located at right angles (90 degrees) to the two major beam lobes produced by each active array.

In operation, these nulls may be used to advantage to eliminate strong interference coming from an undesired direction. The major radiating point on each active element occurs at a point 42 percent up from the bottom end of each vertical. The distance between the bottom horizontal 1/16-wavelength interconnecting wires and earth ground should be no more than 2 to 3 feet for proper performance.

remote switching and balun tap alignment

The simple arrangement of a common 1/4-wavelength stub switching to activate either array provides controlled directivity for both transmitting and receiving. The length of the 1/4-wavelength stub for either band is given in **table 1** as a starting point for the balun tap position.

The tap should be made progressively upward from this overall dimension. Each time the tap location is changed a few inches at a time, the operator should observe an in-line measurement of SWR. The proper tap point will result in the lowest VSWR reading when matching into appropriate coaxial cable using a 1:1 ratio type balun transformer. The use of a balun in this array application is highly recommended to maintain a complete balance of the array system and to minimize beam pattern distortions as well as deterioration of noise pick-up immunity so characteristic of poorly matched or unbalanced feed systems.

The coaxial cable, whose length is not critical, may be buried and its outer shield grounded for further immunity from noise pick-up in severe cases.

Because of weather considerations for rain or snow, the DPDT relay should be mounted inside a waterproof enclosure. This may be combined with supporting the four bottom horizontal 1/16 wavelength wires if the enclosure is elevated by a post at the recommended 2 to 3 feet above the ground level.

ham radio

feeding phased arrays: an alternative method

Carefully chosen feeder lengths provide good match

This article describes a method of feeding a two-element phased-array antenna without using impedance-matching or phase-delay networks at each element. The method requires determining the length of the feedline from each antenna to the "common point" (the point at which the feedlines are combined) and installing a matchbox or antenna tuner at this location (or in the shack) in order to transform the paralleled impedance to 50 ohms.

network method of matching

As was clearly pointed out by Forrest Gehrke, K2BT, in his extensive series of articles on phased arrays, the insertion of a 90-degree delay line into a feeder will, in most cases, *not* guarantee a 90-degree phase shift in the feeder current.[1] If the actual driving-point impedance of each element is known (for a specified current amplitude and phase), then the voltage amplitude and phase at any point on that feeder can be calculated. The technique given by K2BT allows one to choose any convenient length for the antenna feeders and then design suitable networks to alter the feeder voltages so that they are all transformed to the same value of voltage amplitude and phase. These terminals can then be safely and correctly joined together (at the "common point") since they are all at the same voltage. The impedance-matching and phase-delay networks are used to force all the voltages to be the same. Any convenient length of 50-ohm transmission line (the main feeder) can then be utilized to span the distance from the common point to the shack. The main feedline will be "flat" (the SWR will be low) because the networks at the ends of the antenna feeders are designed to present a combined impedance of $50 + j0$ ohms (pure resistance) when "looking" into the common point toward the antennas (see **fig. 1A**).

The method shown in this article calculates the voltages that must exist at various points along the antenna feeders. When a point is found on one antenna feeder where the voltage (both amplitude and phase) is identical to the voltage at some point on another antenna feeder, then those two points can be connected together without altering the relationship of the currents at the driving points of the two antennas. Rarely will the impedance seen looking toward the antennas from this common point be 50 ohms.* Thus, the main feedline running from this point into the shack will not be flat (the SWR will be high) and a matchbox or tuner will be needed at the operating position to present a well-matched load to the transmitter, (see **fig. 1B**). Note: the matchbox could be

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placed into this circuit at the common point if desired. Then the main feeder into the shack would be "flat" as in **fig. 1A**. Of course, the matchbox would then be "dedicated" to the array and could not be used for other purposes unless removed.

calculations are simple

Only one formula, used repeatedly, is needed to calculate the voltages at all points along the antenna feedlines. It's best to utilize a personal computer or programmable calculator to reduce the drudgery of this task.

$$\bar{E}_{IN} = \bar{I}_{OUT}\,(A\bar{Z}_L + B) \qquad \textbf{(1)}$$

where $\bar{E}_{IN}$ = voltage at the input end of the antenna feeder

$\bar{I}_{OUT}$ = current at the output or load end of the antenna feeder
= current at the driving-point of the antenna

$\bar{Z}_L$ = impedance at the output or load end of the feeder
= driving-point impedance of the antenna

A = $\cos\theta$

B = $jZ_0 \sin\theta$

Z_0 = characteristic impedance of the antenna feeder

θ = electrical length of the antenna feeder, in degrees *(360 degrees* = λ)

Note that $\bar{E}_{IN}$, $\bar{I}_{OUT}$, and $\bar{Z}_L$ are all complex numbers; that is, they have both magnitude and phase. The equation therefore requires the use of complex, or vector, arithmetic. *A* and *B* are two of the "ABCD" parameters discussed by K2BT in Part 5 of his series.[2] (In addition, both "*A*" and "*B*" can be complex numbers as well. "*A*" is complex if the cable attenuation is taken into account.)

The goal is to find a place on the antenna No. 1 feeder where the voltage is identical to the voltage somewhere on the antenna No. 2 feeder. To do this, make a list for each antenna, labeling one column for "feeder length in degrees" and the other "voltage at end of feeder." Then calculate and record the voltage on each feeder at 10 degree intervals. After finding 10 or 15 values for each feeder, stop calculating and compare the lists. It is better to record the voltages in polar form (amplitude and phase angle) for easier comparison. Also, keep all the angles less than or equal to ±180 degrees. (For example, an angle of −230 degrees is equivalent to +130 degrees. Always use the smaller number.) When two amplitude values are found that are close to each other, check the angles to see if they are similar. If both amplitude and phase are "in the ballpark," redo the calculation at 1 degree or 1/2 degree increments until two lengths are found where the voltages are identical, or nearly so. If none of the recorded values on one list are comparable to those on the other list, continue the calculations and add more data points to the lists — in other words, make the antenna feeders longer. It should be possible to find the required line lengths within a reasonable amount of time.

*A 1:1 VSWR occurs only when the paralleled driving-point impedance of the elements equals the characteristic impedance of the line. **Editor.**

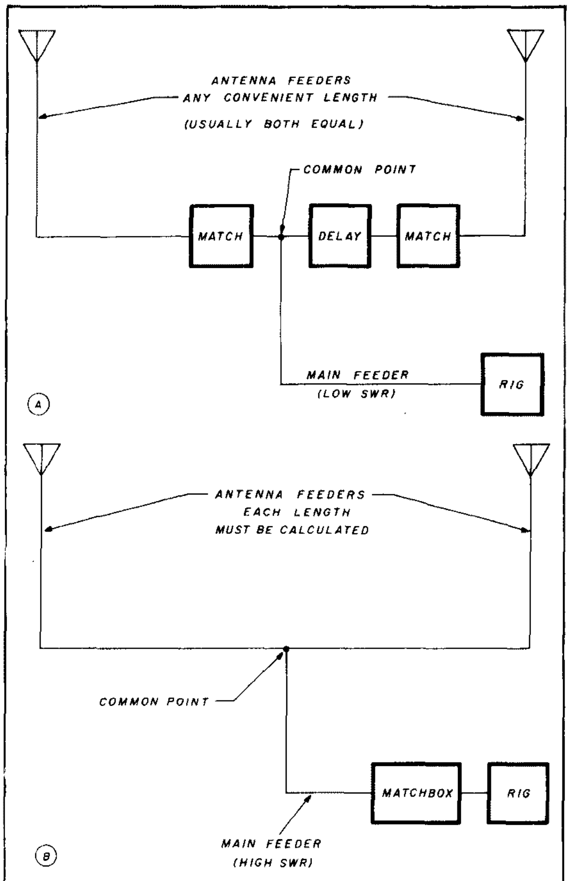

fig. 1. Diagram shows similarities and differences between KB8I's method and that of K2BT. (A) The effect of different feeder lengths are compensated for by the use of individual matching circuits. (B) By moving the matchbox to the common point arbitrary feeder lengths can no longer be used (see text).

cutting the coax

After the required length of each antenna feeder has

been calculated, it's time to actually cut the coax, remembering to take the velocity factor of that particular cable into account. It's a good idea to actually measure the velocity factor of the cable, rather than relying upon the manufacturer's data. Since the electromagnetic fields travel more slowly inside the cable than they do in free space, the physical length of the cable will be shorter than the electrical length:

$$L = 2.734\ (L_e)\ (V_F)/f \qquad (2)$$

where L = physical length of antenna feeder, in feet (1 meter = 3.281 feet)

L_e = electrical length of antenna feeder, in degrees

V_F = velocity factor of the cable, as a decimal (i.e., 0.66)

f = operating frequency, in Megahertz

switching directions

For a two-element array, the pattern may be reversed by interchanging the antenna feeder cables at the antenna terminals. One way to do this, shown in **fig. 2**, is to cut the longer of the two antenna feeders into two pieces, one of which is the same length as the shorter antenna feeder. These two equal-length cables are then connected directly to the antenna feedpoints and the remaining third piece of cable is switched back and forth from one antenna feeder to the other using a DPDT relay. For best results when building a switchable array, the lengths of the radiators and the ground systems under each antenna should be adjusted until both antennas have the same self-impedance.[3]

20-meter array

I became interested in phased arrays while accumulating the contacts necessary for WAS on 20 meters. A home-brew three-element wire beam suspended from trees at 18 feet (5.5 meters) seemed to be performing satisfactorily, and I was working new states at a good pace. However, a phone call revealed that I was also "quite strong" on my neighbor's (just to the west) color TV set. I wasn't too surprised because his portable TV — equipped only with rabbit ears for an antenna — was less than 50 feet (15.2 meters) from my wire beam, and I was running the legal limit. It became vitally important that I construct an antenna that would concentrate all the RF energy toward the east and place a big null right over my neighbor's living room — he is, after all, not only my neighbor, but my landlord.

My backyard was too small for a two-element array with 1/4 wavelength spacing, but the literature revealed that 75 degree spacing (0.208 wavelength) combined with 105-degree phasing would yield the desired cardioid end-fire pattern. Measurements showed that I could squeeze this array into the yard without the radials protruding onto the property of my neighbor to the east. Two tri-band trap verticals were

table 1. Voltages at various points along the antenna feeders.

west antenna ($I = 1\angle 0$ degrees)

θ (degrees)	E_{IN} (volts)
0	15.56 $\angle -45$ degrees
10	11.04 $\angle -11.23$ degrees
20	12.35 $\angle 33.2$ degrees
30	18.17 $\angle 58.4$ degrees
40	25.17 $\angle 70.4$ degrees
50	32.02 $\angle 77.2$ degrees
60	38.2 $\angle 81.7$ degrees
70	43.4 $\angle 85$ degrees
80	47.37 $\angle 87.7$ degrees
90	50 $\angle 90$ degrees
100	51.19 $\angle 92.1$ degrees
110	50.89 $\angle 94.2$ degrees
120	49.11 $\angle 96.4$ degrees
130	45.92 $\angle 98.96$ degrees
140	41.43 $\angle 101.7$ degrees
150	35.82 $\angle 105.4$ degrees
160	29.32 $\angle 110.6$ degrees
170	22.32 $\angle 119$ degrees
180	15.56 $\angle 135$ degrees

east antenna ($I = 1\angle -105$ degrees)

θ (degrees)	E_{IN} (volts)
0	49.68 $\angle -64.9$ degrees
10	54.92 $\angle -57.95$ degrees
20	59.16 $\angle -52.13$ degrees
30	62.14 $\angle -46.98$ degrees
40	63.69 $\angle -42.2$ degrees
50	63.74 $\angle -37.53$ degrees
60	62.27 $\angle -32.77$ degrees
70	59.37 $\angle -27.65$ degrees
80	55.19 $\angle -21.87$ degrees
90	50 $\angle -15$ degrees
100	44.18 $\angle -6.41$ degrees
110	38.31 $\angle 4.83$ degrees
120	33.26 $\angle 19.84$ degrees
130	30.18 $\angle 39.02$ degrees
140	30.09 $\angle 60.32$ degrees
150	33.02 $\angle 79.71$ degrees
160	37.99 $\angle 94.96$ degrees
170	43.84 $\angle 106.39$ degrees
180	49.68 $\angle 115.10$ degrees

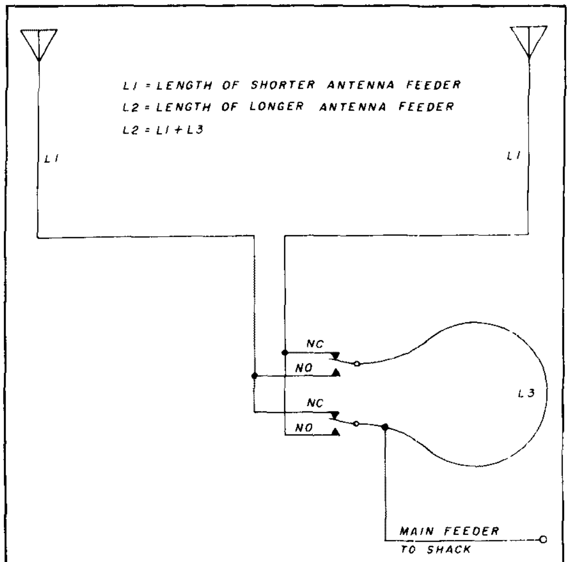

fig. 2. Wiring diagram for DPDT relay used to reverse directivity of array. Directivity is toward the right when connected as shown.

erected, each atop a 15-foot (4.6 meter) mast. They were spaced 75 degrees apart (14.4 feet or 4.4 meters at 14.25 MHz) on an east-west line, and each had twenty 1/4 wavelength radials. The feed system would be designed so that each antenna received the same current amplitude, but the east antenna would lag the west antenna by 105 degrees, placing the null directly on my landlord's living room.

After everything was built, I used an antenna noise bridge and calculator to analyze the system according to the procedure described by K2BT.[3] I measured the self-impedance and driving-point impedance of each vertical at the end of a known length of coax, and then "rotated" each of these values to the antenna feedpoint. This may be done on a Smith chart or through the use of the following formula:

$$Z_L = \frac{D\bar{Z}_{IN} - B}{A - C\bar{Z}_{IN}} \qquad (3)$$

where

Z_L = load impedance at the output end of the transmission line

Z_{IN} = impedance measured at the input end of the transmission line

A = D = cos θ

B = $jZ_0 \sin\theta$

C = $j \sin\theta/Z_0$

Z_0 = characteristic impedance of the transmission line

θ = electrical length of the line, in degrees

While it's better to measure all the impedances right at the antenna feedpoint, I didn't enjoy the prospect of standing on tiptoe perched 10 feet (3 meters) above the ground. I opted for the easier way, and the data obtained enabled me to calculate the mutual impedance between the two verticals. I used this figure to determine the actual driving-point impedance of each element when fed by the currents I had specified for the array. The product of the driving-point impedance and current gives the actual voltage at the feedpoint of each antenna:

west antenna:

$I_1 = 1\,\angle 0 \text{ degrees amperes}$

$Z_1 = 11\text{-}j11 \text{ ohms}$

$= 15.56\,\angle -45 \text{ degrees ohms}$

$E = I_1 Z_1 = 15.56\,\angle -45 \text{ degrees volts}$

east antenna:

$I_2 = 1\,\angle -105 \text{ degrees amperes}$

$Z_2 = 38 + j32 \text{ ohms}$

$= 49.7\,\angle 40.1 \text{ degrees ohms}$

$E_2 = I_2 Z_2 = 49.7\,\angle -64.9 \text{ degrees volts}$

The antenna currents were assumed to have a magnitude of 1 ampere for ease of calculation. The actual current magnitudes will be determined by the amount of drive power. What *is* important is that the actual currents be equal in magnitude, and that the east antenna current lags the west antenna current by 105 degrees.

My lists of the voltages that will occur at various points along each antenna feedline are shown in **table 1**. Scanning the table, we can see that the voltage on the west antenna feeder at a point 50 degrees from the antenna is $32.02\angle 77.2$ degrees volts, while the voltage on the east antenna feeder at a point 150 degrees from the antenna is $33.02\angle 79.71$ degrees volts. These two values are close to each other, and we can now calculate more voltages, this time at 1 degree intervals centered around 50 degrees for the west feeder and 150 degrees for the east feeder. This information is shown in **table 2**. Comparing the voltages on the two feeders, we can see that the values are nearly identical for a 51-degree west feeder ($32.67\angle 77.77$ degrees volts) and a 149-degree east feeder ($32.61\angle 77.95$ degrees volts).

table 2. Voltages at 1 degree intervals along the antenna feeders.

west antenna (I = 1 /0 degrees)

θ (degrees)	E_{IN} (volts)
47	30.02 /75.53 degrees
48	30.69 /76.12 degrees
49	31.36 /76.7 degrees
50	32.02 /77.2 degrees
51	32.67 /77.77 degrees
52	33.32 /78.27 degrees

east antenna (I = 1 / −105 degrees)

θ (degrees)	E_{IN} (volts)
147	31.87 /74.29 degrees
148	32.23 /76.14 degrees
149	32.61 /77.95 degrees
150	33.02 /79.7 degrees
151	33.45 /81.43 degrees
152	33.89 /83.11 degrees

table 3. Theoretical front-to-back ratio of the 2-element phased vertical array.

elevation angle, θ (degrees)	front-to-back ratio (dB)
0	∞
10	39.72
20	27.68
30	20.61
40	15.56
50	11.59
60	8.26
70	5.31
80	2.60

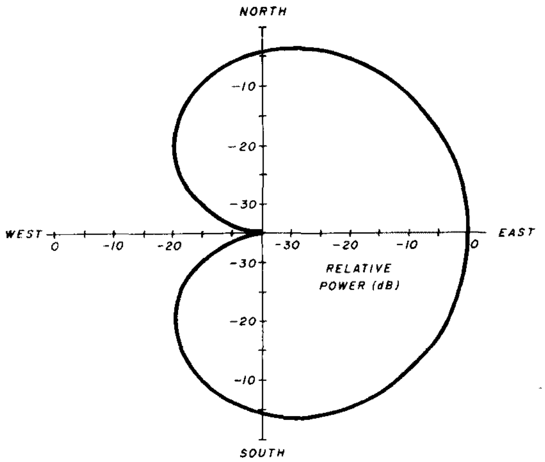

fig. 3. Azimuthal-plane power pattern for two 1/4 wavelength monopoles; spacing = 75 degrees, currents = 1 /0 degree, 1 / −105 degrees, perfect ground; 0 degree elevation angle.

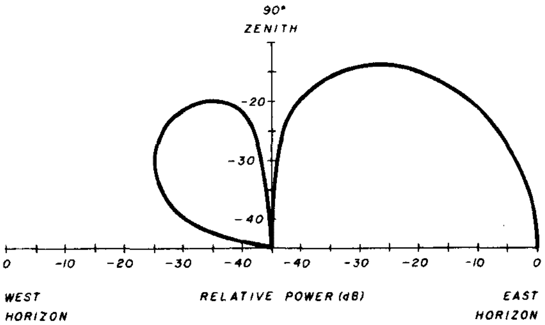

fig. 4. Elevation-plane power pattern for two 1/4 wavelength monopoles; spacing = 75 degrees, currents = 1 /0 degree, 1 / −105 degrees; perfect ground; east-west azimuth.

Notice that the driving-point currents are 105 degrees out of phase, yet the points on the feeders where the voltages are in phase are only (149 degrees − 51 degrees =) 98 degrees apart. Also notice that the east feeder, which has the lagging feedpoint current, is longer than the west feeder, which one would expect.

Measurements on the RG8X cable used for the antenna feeders showed that its actual velocity factor was 0.725 rather than the advertised value of 0.81. Using **eq. 2**, the actual lengths of coax needed were 7.094 feet (2.162 meters) and 20.726 feet (6.317 meters). These two antenna feeders were cut to length and installed. A piece of RG-8 about 30 feet (9.1 meters) long extended from the Tee connector, where the antenna feeders were joined, to the matchbox in the shack.

The array performed as expected. State number 50 (Delaware) was worked shortly thereafter, along with contacts in Africa. There were almost no contacts to the west — but there also were no further TVI problems. The major azimuthal and elevation-plane radiation patterns are shown in **figs. 3** and **4**; the front-to-back ratio in the main elevation plane is provided in **table 3**. Notice that the front-to-back ratio is infinite only on the horizon (0 degree elevation angle) and deteriorates to just one or two S-units at high angles. There is actually a fairly large rear lobe whose maximum occurs at an elevation angle of about 60 degrees. Thus, there will be a fair amount of radiation and signal pickup off the back of the array if propagation favors this high angle.

conclusion

A technique has been demonstrated for feeding two-element phased arrays based upon the premise that the feeders from individual radiating elements may be directly connected together if the voltages at the point of connection are identical in magnitude and phase. A method was shown for determining the voltage at any point on an antenna feeder when the impedance and current at the driving-point of the antenna are given. Using these ideas, an actual 2-element array can be built and operated successfully on the air.

acknowledgements

The author would like to thank Forrest Gehrke, K2BT, for his encouragement, especially during the early stages of this project in 1983. Appreciation is also expressed to Margaret Shields in the Department of Electrical and Computer Engineering at Ohio University for typing the manuscript.

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ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

the first Radio Amateur

The world in 1900 was a far different place than the world we know. No one had heard of credit cards or ball point pens; time sharing meant togetherness, not computers. There were no fluorescent lights, instant coffee, antibiotics, Frisbees or frozen foods. There was no television, no radio — and no Amateur Radio.

Into this late Victorian era there stepped a Giant. In 1896 a young Italian arrived in England with some mysterious scientific apparatus in his luggage. The Customs officials, who had seen nothing like it before, examined it so thoroughly that the delicate apparatus was completely wrecked. This was the inauspicious beginning of a venture destined to remold the pattern of 20th-century living.

Guglielmo Marconi, at 26, had come to England to seek aid in developing his latest invention, a means of signalling at a distance without wires. The hub of a great Empire, Britain possessed the world's greatest mercantile fleet and the mightiest Navy — and it was in shipping that Marconi could see his dream come true. For once a ship left the sight of land it was isolated from the world. When disaster struck — as it often did — some form of communication between ship and shore was sorely needed.

Marconi demonstrated his equipment to the War Office and the British Post Office. The War Office was interested, but the idea of an alternative means of communication seemed to be unpleasant to the Post Office; it was unenthusiastic.

By 1898 Marconi had successfully demonstrated overland and ship-to-shore communication and had formed the Wireless Telegraph and Signal Company Limited. From time to time, communication across distances of up to 100 miles could be established. Yet therein lay an enigma. Hertz's early experiments showed that the invisible "wireless" waves obeyed the laws of light and travelled in straight lines. How, then, could Marconi communicate beyond the visible horizon? The scientific community regarded Marconi's experiments with caution.

The Post Office monopoly on long distance telegraph communication and other restrictions made it impossible for Marconi to set up a revenue-earning inland wireless telegraph service. His company was rapidly going bankrupt. What to do?

the transatlantic gamble

Marconi had two ambitions: one, to prove his system offered dependable long-range capability and two, to compete directly with the Post Office, breaking its monopoly, and maintaining a profitable wireless telegraph service. He conceived a bold stunt to draw attention to his plan: he would send a wireless signal across the Atlantic! Marconi proposed a station of breathtaking power, size, cost, and complexity. It was like proposing to build a cathedral in a world which had seen *nothing more* grandiose than a log hut. The directors of his company objected to taking such a risk. His grand idea was met with scorn and disdain.

But Marconi convinced his colleagues to go along with his plan. The station, to be built in Poldhu, Wales, was staggering in concept; a 25 kilowatt power plant would drive a two-stage spark transmitter (**fig. 1**) connected to a 400-wire antenna supported by 20 masts, each 200 feet high.

By early 1901 the station began to take shape. Preliminary tests indicated a range of at least 225 miles. Despite a myriad of troubles, the station could be heard as far away as Ireland. It had already proven that wireless waves followed the curvature of the globe; Marconi was certain that his signal could be extended to North America.

In March, 1901, Marconi sailed for America and chose the site for a second station: South Wellfleet, a small town on the eastern shore of Cape Cod, Massachusetts. Leaving his chief engineer there, Marconi returned to England.

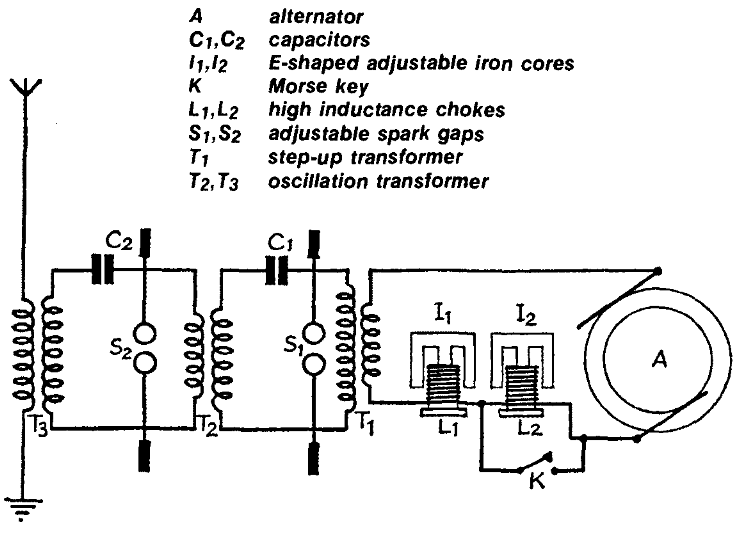

fig. 1. **Marconi spark transmitter at Poldhu.**

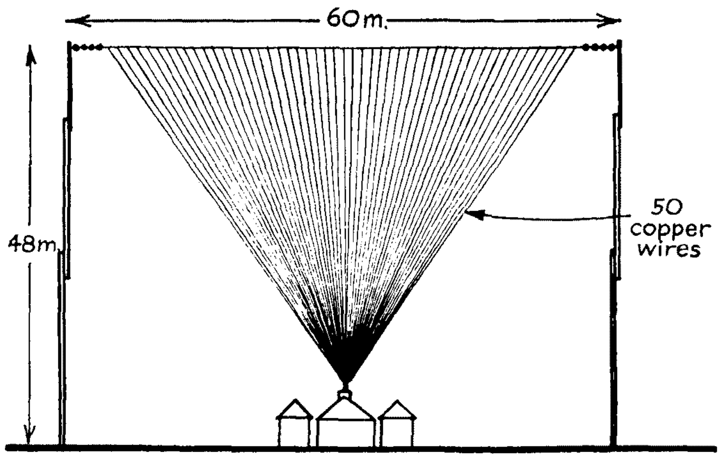

fig. 2. The temporary antenna used at Poldhu for the transatlantic tests.

Catastrophe struck in September when a vicious gale hit the coast of Wales, breaking one of the many guy wires. All 20 masts collapsed into a shambles of broken timber and tangled wire. Just a month later a second storm destroyed the antenna system at South Wellfleet.

The directors of the Marconi Company were appalled. Over a quarter million dollars had been spent with nothing to show for it but chaos. Marconi would not give up; he cleared away the wreckage, erected a temporary antenna (**fig.** 2) and had the station back on the air in 11 days.

Because of the diminished capacity of the makeshift station, Marconi decided to abandon the Wellfleet site and set up temporary receiving equipment in Newfoundland, the point of landfall nearest to Wales. In great secrecy he set sail for St. John, with a small stock of kites and balloons to keep a single wire aloft in the stormy weather.

A site was chosen on Signal Hill, and on December 9 the apparatus was assembled in an abandoned military hospital. The balloon was prepared for inflation and ground plates were buried. A cable was sent to Poldhu requesting that the Morse letter "S" be transmitted continuously from 3 to 7 PM. Marconi chose his message wisely. He knew the fragile state of his equipment, and that the transmission of dashes, rather than dots, would have imposed too great a strain on the keyer and the transformers.

On December 10th the weather was fair. A balloon-supported kite antenna was sent aloft. The transmissions started at a power level of about 10 kilowatts and on a wavelength of about 366 meters (820 kHz). Since there was no means of measuring frequency, the actual wavelength remained a matter of speculation.

As the wind picked up, the balloon bobbed and weaved about in the sky above Signal Hill. Marconi adjusted his new "syntonic receiver" — a glass tube within which a globule of mercury was held between two iron or carbon rods, forming a crude semiconductor. Nothing which could be identified as the letter "S" could be heard amid the static. The wind picked up and the antenna crashed to earth as the balloon was swept away.

December 12, 1901

On the 12th the wind increased in intensity. A kite was launched bearing a 510-foot wire. The wind carried it away. A second kite was launched with a 500-foot wire attached. Because he had observed that the buffeting of the antenna by the wind made it impossible to keep the newer receiver in tune, Marconi sat listening intently at an older, untuned receiver (**fig. 3**). Time slipped by. Suddenly, at 12:30 PM, Newfoundland time, he handed the earphone to George Kemp, his

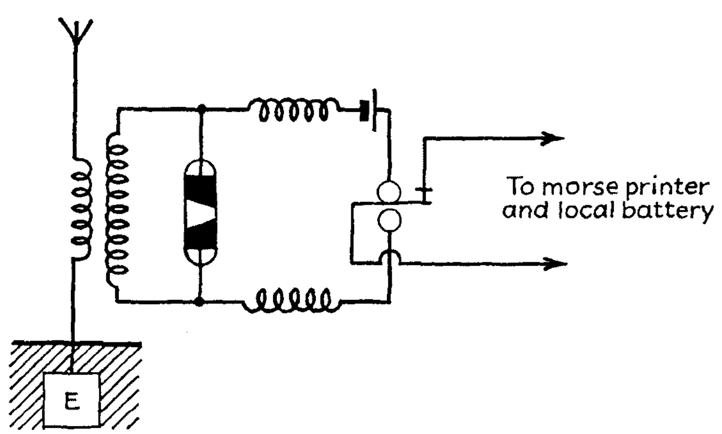

fig. 3. Marconi's untuned receiver.

assistant, and quietly asked, "Can you hear anything, Mr. Kemp?"

George Kemp took the headphone. Through the static crashes he could hear, faintly, the unmistakable rhythm of three clicks, followed by a pause, then three more and a pause, and so on, until — all too soon — the signals were lost once more in static. Marconi, a cool-headed man if there ever was one, wrote in his laboratory notebook: *Sigs at 12.20, 1.10 and 2.20.*

Marconi a fraud?

Marconi was in a quandry. What conclusive proof did he have? He and Kemp were not exactly unbiased. Should he make a public statement?

Finally on December 14th Marconi cabled his company the news. It was made public on December 16th, 1901.

The first reaction came from the lawyers of the Anglo-American Telegraph Company, whose cable line had carried the message to England two days previously. It was sharp and to the point. Marconi was told that the company had a monopoly on communication in Newfoundland and it forbade any future infringement of their rights under pain of legal action.

The public interest, however, was aroused and both the Canadian and American governments expressed interest in the experiment. The technical journals treated the incident with a combination of skepticism and indignation. Marconi had no proof to substantiate his claim, which challenged the fundamental laws of physics and the proven knowledge of Newton, Maxwell, Hertz, and others. It was not until later, when the reception of signals across the Atlantic was demonstrated beyond any shadow of doubt, that Marconi's achievement was recognized.

Even today, it is difficult to believe that the 366-meter signals could actually have been heard. The receiving equipment after all, consisted of an inefficient antenna coupled to an untuned receiver which had no means of amplification whatsoever and was even less sensitive than crystal detectors, which evolved a few years later. If, in fact, the wavelength was 366 meters, the tests took place at the worst possible time of day because the entire path would have been in daylight. Today we know that radio signals can travel across the Atlantic and far beyond. But in 1901, anyone who believed that they could, and did, believed so as an act of faith based upon the integrity of one man — Marconi.

Marconi at the World's Fair

It was 1932. Marconi, an internationally acclaimed scientist, inventor, and businessman was in the United States. He was scheduled to make an official visit to the Chicago World's Fair, a breathtaking exhibition of the modern world of technology. Fair officials were in a dither as Marconi arrived in the company of other important dignitaries. The Great Man toured the Fair, expressing great interest in the scientific exhibits. News photographers crowded around Marconi and he was followed by a large gathering, all craning their necks to see the Father of radio communication.

As he was about to leave, Marconi expressed a wish to visit the Amateur Radio station at the Fair. So, Marconi's big, black limousine, with colorful American and Italian flags flying from the fenders, drew up in front of a building on the edge of the fairground, followed by a horde of officials and newsmen. Marconi was escorted up the stairs to the World's Fair Amateur station, W9USA.

The young operators of W9USA appeared thunderstruck as the famous visitor strode into the station, introduced himself, and studied the homemade kilowatt trasmitter and the superheterodyne receiver. He examined the station log book. The Fair officials were mystified by the incomprehensible collection of equipment that seemed to fascinate Marconi.

One of the operators apologized to Marconi, saying that the equipment had been built by mere Radio Amateurs. Marconi nodded and smiled, shaking the hand of the operator warmly. "Yes, yes," he said. "I understand — after all, I am a Radio Amateur myself."

radio silence

Marconi died in July, 1937. On the evening of the following day, at the state funeral in Italy, the Italian Radio Service observed an official period of radio silence. In England, and throughout the world, thousands of radio stations — broadcast, commercial, and Amateur — fell silent. The radio silence which Marconi had broken when he switched on his first transmitter came down in sorrow at his passing.

Shortly before he died, Marconi, in an address at St. Andrews University in Edinburgh, Scotland, stressed his original intention to make the high seas safe by giving ships a means of communication. Then, perhaps talking more to himself than to his audience, he added, "Have I done the world good — or have I added a menace?"

in retrospect

Much speculation has taken place during the decades following Marconi's famous transatlantic experiment. Ionospheric studies and a review of the sunspot cycle suggest that propagation across the Atlantic at that time of day, in that month and year, was highly unlikely on 820 kHz.

What, then, might Marconi and his assistant have heard? A few clues exist. Marconi's spark transmitter emitted a rough wave, high in harmonic content. He used a broadband (untuned) receiver. Perhaps Marconi was not hearing the fundamental transmitter signal, but instead a harmonic of the signal, in particular the fourth harmonic on about 3280 kHz. If that is so, then a few hundred watts of harmonic power would have easily made the transatlantic journey.

Of course, with a broadband receiver, Marconi could have heard many harmonics at the same time! So we shall never really be sure what transpired on that stormy day in Newfoundland. It is interesting, however, to think that if Marconi had elected to listen to his tuned receiver, used in the previous day's unsuccessful tests, he might have heard nothing at all!

acknowledgements

Thanks to the Official Historian of the Marconi Company (England) for the background information on Marconi's early experiments. The story of Marconi's visit to the Chicago World's Fair appeared in the November, 1932 issue of *QST*.

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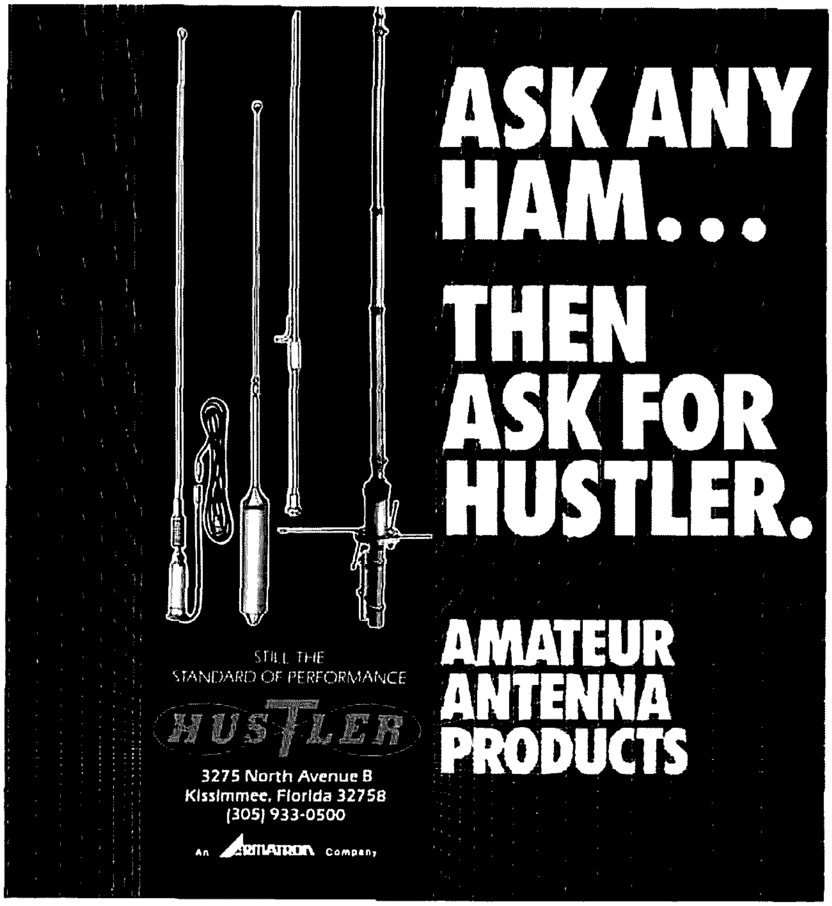
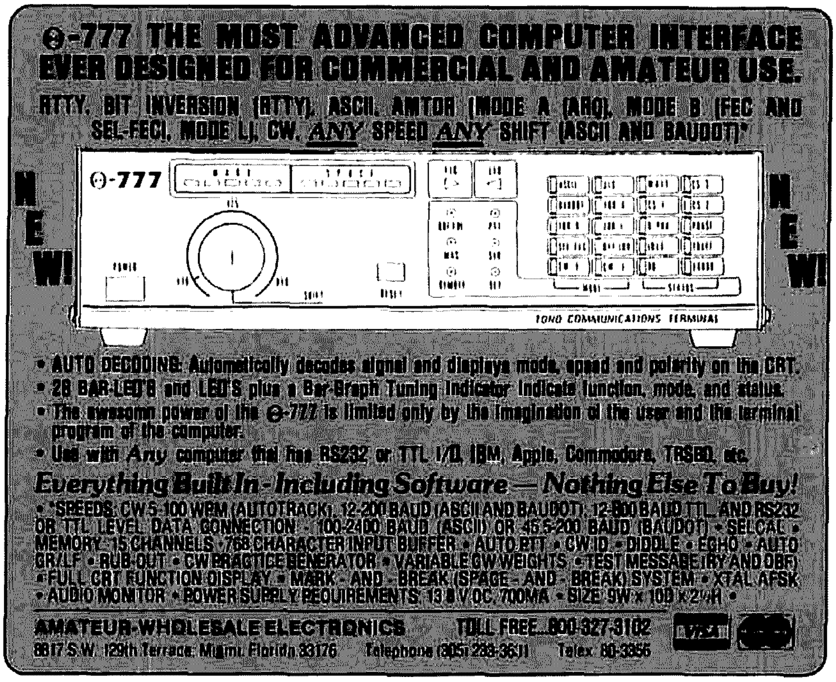
*Drawings adapted from *A History of the Marconi Company* by W.J. Baker, Methuen & Co., Ltd., London, 1970. (Distributed by the Marconi Co., Ltd.)

ham radio



141



140

a sloping terminated vee beam

Low SWR across ham and SWL bands

This is the antenna that every ham has probably dreamed about: a beam antenna that works on every Amateur band and on all the spaces in between . . . an antenna that's cheap and requires just a single tree or pole for support . . . an antenna that doesn't require precision measurements, special tuning instruments, or almost anything else out of the ordinary. *Everything* needed can be purchased inexpensively or built at home, and the hardest part will be building a toroid transformer with about 25 turns of wire.

This is the same antenna used as a back-up by the Voice of America, so you know it works.

According to transmission line theory, a single conductor placed a few feet above and parallel to the ground can be used as an antenna. If the wire is raised more than a few feet above ground, the impedance approaches a value of 400 ohms according to the following formula:

$$Z_0 = 83.7 \log_{10}(4H/d) \tag{1}$$

where Z_0 = impedance of single wire above ground
H = height of the wire above ground in any units
d = diameter of the wire in the same units

The nonresonant "vee" antenna and a large family of similar antennas such as rhombics are based on transmission line theory. When used as an antenna, however, the transmission line has spacing which is large compared with normal lines. As a result it is leaky and can be used as an antenna. The radiation field from a single wire transmission line may be calculated[1] and the two maximum gain lobes discovered by the equation:

$$E(a) = \frac{60 \sin a \bullet \sin\left[\frac{\pi L}{W}(1-\cos a)\right]}{1 - \cos a} \tag{2}$$

where $E(a)$ = normalized field strength at angle a
a = the angle from the axis of the wire
π = 3.1416
L = length of the wire in any units
W = one full wavelength in the same units

If two terminated transmission lines are aligned so that the major lobes of each wire are additive, the lines then become a "vee" antenna. (Two vees placed end-to-end form a rhombic.) Because the line is terminated in a matching impedance, the wires are nonresonant — just like any correctly matched, well-behaved transmission line — and the antenna will accept power at any frequency.

Several good things happen with terminated transmission line antennas. One desirable result is broadbanding operation; another is that the reverse lobes (toward the pointed side of the vee) are cancelled or greatly reduced. The resulting antenna can be used to both transmit and receive at every frequency within its range, much like a log periodic antenna. But, it will

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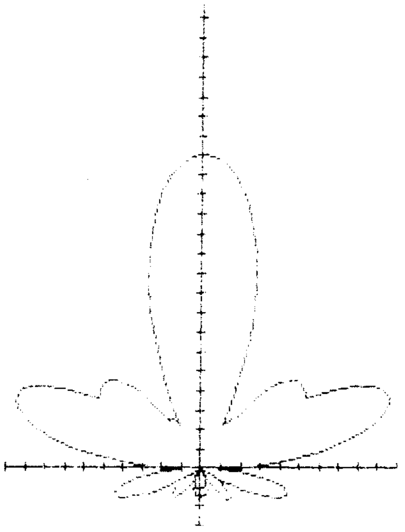

fig. 1. A maximum lobe gain of 5.1 dBi is achieved with a 1-wavelength leg length 68-degree apex angle, non-resonant vee beam.

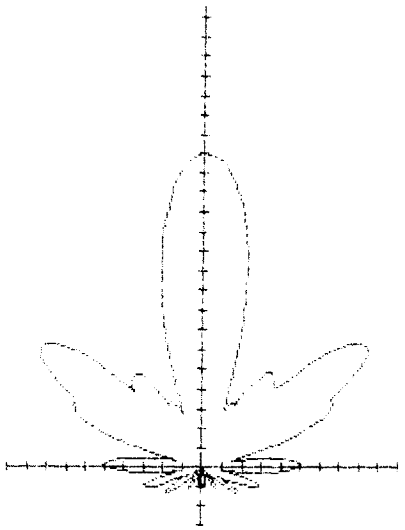

fig. 2. A maximum lobe gain of 5.7 dBi is achieved with a 1.5-wavelength leg length 56-degree apex angle, non-resonant vee beam.

table 1. Relationship between vee beam leg lengths, apex angles, and gains.

vee leg length in half waves	apex angle in degrees	gain in dBi
2	68	5.1
3	56	5.7
4	48	6.1
5	44	6.4
10	32	7.4
20	22	8.4

produce its best gain only over a relatively narrow frequency range where the major lobes align for best gain. This frequency range depends on the angle of the apex of the vee and the length of the legs. For many purposes, +100 percent to −50 percent frequency excursion is reasonable, (see **table 1**). The worst thing that happens is that the beam splits and you find a null in the center with the gain lobes on either side.

In this case the vee has been optimized for the 14 to 15 MHz band — but it also works reasonably well from 40 to 10 and even radiates modestly on 160 and 80 meters. Optimization can be done for any band by calculating the major lobes and adjusting the apex angle for that band. Ordinarily, however, not much is gained by changing the apex angle of such a short vee. **Figures 1** through **6** illustrate the azimuthal pattern of vee beams with leg lengths from 1 to 10 wavelengths.

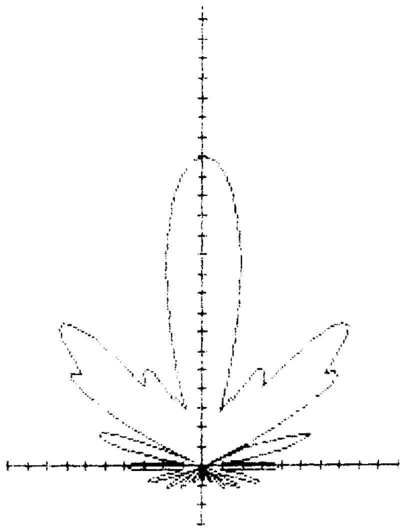

fig. 3. A maximum lobe gain of 6.1 dBi is achieved with a 2-wavelength leg length 48-degree apex angle, non-resonant vee beam.

construction is easy

The vee is probably the simplest of all wide band beam antennas to build. A single support roughly 65 feet (20 meters) tall is required for the high end and two supports about 3 feet (1 meter) tall for the low ends. The only critical item is the pointing of the

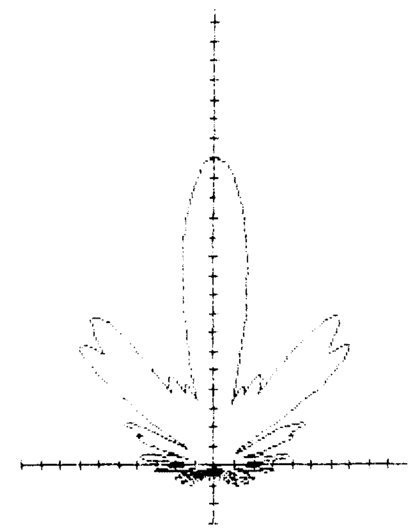

fig. 4. A maximum lobe gain of 6.4 dBi is achieved with a 2.5-wavelength leg length, 44-degree apex angle, nonresonant vee beam.

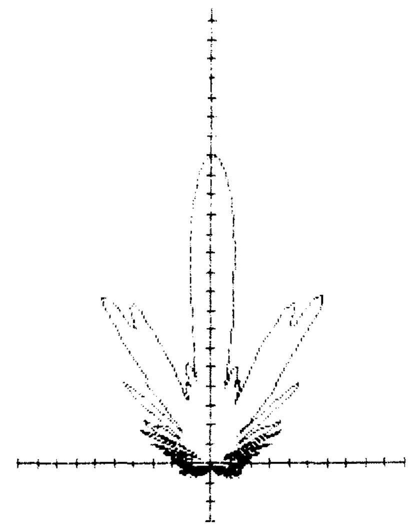

fig. 5. A maximum lobe gain of 7.4 dBi is achieved with a 5-wavelength leg length 32-degree apex angle, nonresonant vee beam.

antenna. The position of the two low point anchor poles must be selected to direct the antenna at your target by using a compass bearing corrected for local magnetic variations. In Washington, D.C., the variation is 8 degrees west; thus 8 degrees must be added to the compass reading to get a true bearing. The legs must be located so that they are 22 degrees on either side of the great circle line of bearing to your target and to make the open end of the vee point toward the target. For example, on my globe London and Europe are 48 true (not magnetic) degrees from Washington, D.C.

materials are inexpensive

The wire and insulators can be of almost any kind. However, I prefer No. 14 AWG copperweld wire and No. 500 strain insulators (also known as guy wire insulators). Strain insulators are particularly good if you use a tree for your antenna pole because they won't break when the tree sways in the wind. I also advise soldering all antenna connections to avoid long-term problems from corrosion and the resulting bad connections.

The transformer can be purchased from at least one company for about $150,[2] but building your own transformer is not difficult. Only a few turns of wire on a 2-inch diameter RF toroid core will do the job.[3] Generally the largest RF toroid core which the manufacturer says will cover the desired frequency range will work reasonably well. On my transformer I used 20 turns of soft No. 12 AWG enameled copper wire for the secondary, spacing it evenly around the core and making sure that there were at least a few millimeters between the two wire ends to prevent RF voltage flash-over. If you leave the ends of the secondary about two feet long you can also use them for antenna connections later. Anchoring the wire in place with silicone rubber is a messy but a very effective measure. Then slip

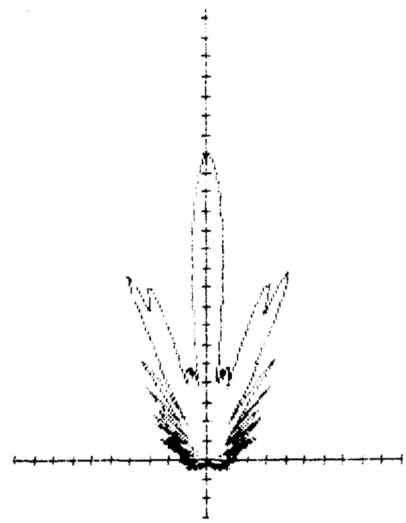

fig. 6. A maximum lobe gain of 8.4 dBi is achieved with a 10-wavelength leg length, 22-degree apex angle, nonresonant vee beam.

some tube insulation over a second piece of the No. 12 AWG enameled wire and wrap five primary turns evenly over the outside of the first twenty turns. Getting even a semblance of a smooth wrap will be the hardest thing you'll have to do for this antenna. Again, use some silicone rubber to tack down the primary and give it at least twenty-four hours to harden. After stripping the enamel from the ends of toroid primary wire, solder on the coax connector of your choice. Resist the urge to use the hole through the middle of the toroid to hold up the high end of the antenna or attach the coax — powdered iron cores are *brittle* and are subject to breaking at the most inconvenient times.

The 390-ohm termination resistors, are noninductive resistors made specially for terminating antennas, and are available from electronic wholesalers,[4] at ham fests (your best chance) or at surplus houses. On the other hand, you can also make a 200-watt terminating resistor out of one-hundred, 2-watt, 39 kilohm, carbon composition resistors. A simple way to do this is to lay out two tinned No. 12 AWG wires spaced the width of the resistor body and solder the 100 resistors to them.

Operating this antenna is pure joy. It loads my transmitter well from one end of the short wave bands to the other. My antenna also shows a maximum of about 1.5:1 SWR on 15 meters and about 1.2:1 maximum on all other bands. Adding or subtracting a turn or two on the transformer optimizes the SWR for any desired frequency.

At last I have an antenna that works smoothly from 160 to 10 meters and even works as an effective all-band antenna for my general coverage short-wave receiver. For cost and coverage, it's hard to beat this *nearly foolproof* sloping vee beam.

references

1. Landee, Davis & Albrecht, *Electronic Designers Handbook*, McGraw-Hill, New York, 1957, (see section 21).

2. The vee antenna feeding transformer, 1.8 to 30 MHz, is available from Apcom Inc., 625 Loffstrand Lane, Rockville, Maryland 20850, (301-294-9060). Note: this company and others will supply the complete antenna, mast, ground rods, earth anchors and everything else, optimized for your frequency and adjusted to fit various larger pieces of land. This is a good deal for commercial stations, but the cost is very high by ham standards.

3. Type F240-Q1 toroid core to cover 1.8 to 30 MHz is available from Amidon Associates, 12033 Otsego Street, North Hollywood, California 91607, or Palomar Engineers 1924-F W. Mission Road, Escondido, California 92025.

4. Carborundum Co., Graphite Products Division, Box 339, Niagara Falls, New York 14302. Please note that this is a manufacturer, not a retail outlet. If you can put together a large order, then request noninductive resistor type 888SP, 390 ohms (2 for each antenna). If more power is required, order type 891SP, 450 ohms (2 for each antenna).

ham radio

DX FORECASTER

Garth Stonehocker, KØRYW

what's this Sporadic E?

In these **minimum** years of the 10.7-year sunspot cycle when the 10-meter band is at best open only a few hours per day and only to the southern directions for long skip, it's still possible for propagation modes to exist on 6 meters and occasionally even the 2-meter band for short-skip openings. Signals appear suddenly, out of nowhere, and frequently rise to amazing strength. They may stay in for just a few minutes at a time — or the band may remain open for hours. Occasionally in June or July DX signals might be heard around the clock. Signals can be received from distances of 500 to 1200 miles and occasionally, due to multiple reflections, from distances as far away as 2500 miles.

How do you recognize such E_s openings? Say you're monitoring a beacon. The band is quiet. Suddenly you hear a buildup of "antenna noise." Almost instantly there are DX stations all over the band. Signal levels fluctuate rapidly as the session opens and as it declines. When the signal is there it usually pegs your S-meter, but it is also subject to rapid fades on the order of 60 dB or more that chop it into a garbled mess.

One way to recognize the probable opening of E_s on 6 meters was reported by George Jacobs, W3ASK, in the June, 1962, issue of *CQ*. Working a lower frequency band, say 15 or 10 meters, listen to the stations being worked. If the minimum skip distance is decreasing, the skywave geometry is such that the maximum usable frequency will be increasing by reflection from an E_s (more dense than F2 and lower height) cloud. W3ASK's rule of thumb states that when stations are heard less than 500 miles away on 10 meters, or less than 350 miles on 15 meters, the chances are good that 6 meters will open in that same direction. It's worth a try! (More on E_s DXing next month.)

last-minute forecast

DX conditions on the higher HF bands, 10 through 30 meters, are expected to be very good the first two weeks of the month and also again during the last week of this five-week month. The propagation at these times can be via long-skip if the radio flux is very high — above about 85 units. Sporadic E (E_s) short-skip is expected to occur occasionally, with the probability of occurrence increasing as the month goes by. The middle two weeks of the month will offer the best nightime DXing. Geomagnetic disturbances are expected to occur around the 4th through the 8th and again on or about the 15th. Concurrently, the thunderstorm noise background level will also be higher on the lower bands.

Moonbounce DXers can take advantage of the lunar perigee (and full moon) that will occur on the 3rd and 4th of this month. An Aquarid meteor shower of interest to meteor-scatter and meteor-burst DXers peaks between May 4th and 6th with rates of 10 and 25 per hour for the Northern and Southern Hemisphere, respectively.

Two eclipse events are calculated to occur in May. The first is a total eclipse of the moon on May 4 from 1817 to 2135 UTC. The path of its shadow starts near New Zealand and travels west through Australia, Asia, Africa into the Atlantic Ocean. The width extends from the Antarctic to Europe. The second event, partial eclipse of the sun, (with 84 percent coverage) will occur on May 19 from 1925 to 2342 UTC. Its path begins at Greenland, travels across Canada and then moves on to Alaska and Japan. The second eclipse provides an opportunity to evaluate propagation effects on the path between the United States and Europe. Schedule contacts for the day before the day of the eclipse and the day after. Compare signal strengths and quality on *each band used* (suggestion: try 40, 30, 20 and 15 meters). A club-coordinated effort is a good way to cover the bands needed.

band-by-band summary

Six meters may provide an occasional opening to South Africa and South America by short-skip E_s. These will occur around local noontime, toward the end of the month.

Ten and fifteen meters will provide a few short skip E_s openings and many long-skip openings, especially during high solar flux conditions, to most areas of the world during daylight. Some transequatorial (TE) openings, associated with a mildly disturbed geomagnetic field may occur in the evening hours. This is about the last month that affords many good TE openings until next fall.

Twenty, thirty, and forty meters will support DX propagation from most areas of the world during the daytime and into the evening hours almost every day. Forty meters has joined this daytime DX group because of lower signal absorption, lower LUF (lowest usable frequency) during these sunspot minimum years. During unusually high 27-day solar flux days 40 meters may not be usable and both 30 and 40 meters may not be usable in the predawn hours after a high flux day. The DX on these bands may be either long-skip to 2500 miles (4000 km) or short-skip E_s to 1250 miles (2000 km) per hop. The length of daylight is approaching maximum, providing many hours of good DXing.

Thirty, forty, eighty, and one-sixty are all good for nighttime DX. Although the background thunderstorm noise is beginning to be noticeable these bands are still quiet enough to provide good DX working conditions. Sproadic-E propagation may be a contributing factor toward enhanced conditions at local sunset and will occur more often during the next three months.

MAY

WESTERN USA

GMT	PDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	5:00	20	20	15	15	20	10	15*	20
0100	6:00	20	20	20	10	20	10	10	20
0200	7:00	20	20	20	10	20	10	10	20
0300	8:00	20	20	20	10	20	10	10	20
0400	9:00	20	20	20	10	20	15	10	20
0500	10:00	20	20	20	10	20	15	15	20
0600	11:00	20	20	20	15	30	15	15	20
0700	12:00	20	30	20	15	*30*	20	15	20
0800	1:00	20	30	20	15	*30*	20	*20*	20
0900	2:00	20	*30*	20	20	*30*	*20*	*20*	20
1000	3:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*20*
1100	4:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*30*
1200	5:00	*20*	20	15	*20*	*30*	*20*	*20*	*30*
1300	6:00	*20*	20	15	*20*	*30*	*20*	*20*	*30*
1400	7:00	*20*	20	15	*20*	*30*	*20*	*20*	*30*
1500	8:00	*20*	20	15	*20*	20	*20*	*20*	*30*
1600	9:00	20	20	15	*20*	20	15	*20*	*30*
1700	10:00	20	20	15	15	20	15	20	20
1800	11:00	20	20	10	15	20	15	20	20
1900	12:00	20	20	10	15	20	15	15	20
2000	1:00	20	20	10	15	20	15	15	20
2100	2:00	20	20	10	15	20	10	15	20
2200	3:00	20	20	10	15	20	10	15	20
2300	4:00	20	20	15	15	20	10	15	20

MID USA

GMT	MDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)	CDT
0000	6:00	20	20	15	15	20	10	10	20	7:00
0100	7:00	20	20	20	20	20	10	10	20	8:00
0200	8:00	20	20	20	20	20	10	10	20	9:00
0300	9:00	20	20	20	20	20	10	10	20	10:00
0400	10:00	20	20	20	20	30	15	10	20	11:00
0500	11:00	20	20	20	20	*30*	15	15	20	12:00
0600	12:00	*20*	20	20	20	*30*	15	15	20	1:00
0700	1:00	*20*	20	20	20	*30*	20	15	20	2:00
0800	2:00	*20*	*30*	20	*20*	*30*	20	*20*	20	3:00
0900	3:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	20	4:00
1000	4:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	*20*	5:00
1100	5:00	20	*30*	15	15	*30*	*20*	*20*	*20*	6:00
1200	6:00	20	20	15	15	*30*	*20*	*20*	*20*	7:00
1300	7:00	20	20	15	15	*30*	*20*	*20*	*20*	8:00
1400	8:00	20	20	15	15	*30*	*20*	*20*	*20*	9:00
1500	9:00	20	20	15	15	*30*	*20*	*20*	*20*	10:00
1600	10:00	20	20	15	15	20	15	*20*	*20*	11:00
1700	11:00	20	20	15	10	20	15	20	20	12:00
1800	12:00	20	20	15	10	20	15	15	20	1:00
1900	1:00	20	20	10	10	20	15	15	20	2:00
2000	2:00	20	20	10	10	20	15	15	20	3:00
2100	3:00	20	20	10	10	20	10	15	20	4:00
2200	4:00	20	20	15	15	20	10	15	20	5:00
2300	5:00	20	20	15	15	20	10	15	20	6:00

EASTERN USA

GMT	EDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	8:00	20	20	15	15	20	10	15	20
0100	9:00	20	20	20	20	20	10	10	20
0200	10:00	20	20	20	20	20	10	10	20
0300	11:00	20	20	20	20	20	10	10	20
0400	12:00	20	20	20	20	30	15	10	20
0500	1:00	20	20	20	20	30	15	15	20
0600	2:00	*20*	30	20	20	*30*	15	15	20
0700	3:00	*20*	*30*	20	20	*30*	20	15	20
0800	4:00	*20*	*30*	20	20	*30*	20	*20*	20
0900	5:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	20
1000	6:00	*20*	*30*	*20*	*20*	*30*	*20*	*20*	20
1100	7:00	20	*30*	15	15	*30*	*20*	*20*	*20*
1200	8:00	20	20	15	15	*30*	*20*	*20*	*20*
1300	9:00	20	20	15	15	*30*	*20*	*20*	*20*
1400	10:00	20	20	15	15	*30*	*20*	*20*	*20*
1500	11:00	20	20	15	15	30	20	*20*	*20*
1600	12:00	20	20	15	15	20	15	20	*20*
1700	1:00	20	20	15	10	20	15	20	20
1800	2:00	20	20	10	10	20	15	15	20
1900	3:00	20	20	10	10	20	15	15	20
2000	4:00	20	20	10	10	20	15	15	20
2100	5:00	20	20	10	10	20	10	15	20
2200	6:00	20	20	15	15	20	10	15	20
2300	7:00	20	20	15	15	20	10	15	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

160-meter transmission line antenna

If height or space is a problem, try this

If you're like me, you don't have adequate yard space to put up a half-wave dipole on the 160-meter band or a tower to load as a short vertical. In the past, I tried not to let this discourage me from getting on 160, but the RF burns and unanswered calls that resulted from loading up a 40-meter dipole forced me to come up with a better antenna!

Short transmission line antennas have been used at UHF and microwave frequencies for quite some time.[1] Small slots carved into the bodies of fast-moving vehicles (airplanes and rockets) have been used to effectively radiate RF energy using transmission line principles. In fact the folded dipole, commonly used with FM receivers, uses these same principles to receive RF signals. Other types of transmission line antennas include the "low-profile" type used on trains and emergency vehicles, where the antenna structure protrudes just fractions of a wavelength above the vehicle body. These antennas are advantageous when antenna size and height are extremely limited.

I live in an apartment complex. The tallest structures are a couple of 30-foot trees in my small backyard. I have also found that good grounding is a problem, making the use of an "RF-free" tuner difficult. After attempting to work the top band with my existing 20-meter and 40-meter dipoles (none too successfully), I decided to give these transmission line ideas a try. The results have been gratifying, to say the least. I now have a resonant 160-meter antenna that requires no tuner, is directly coax-fed, has given no trace of RF in the shack, and provides enhanced reception due to very low noise pick-up.

how it works

Many people associate the term "transmission line" with the coaxial cable or ladder line that feeds their antenna — something that "carries power to the antenna," and not something that should, itself, radiate RF. Of course, it is undesirable to have our feedline radiate, but many successful antennas, such as the longwire, the rhombic, and the Beverage are indeed unbalanced (radiating) transmission line extensions of their feed systems. By configuring these lines properly, resulting current distributions along the wires enable these transmission line extensions to emit and receive far-field RF energy. By analyzing a familiar transmission line antenna, the half-wave folded dipole, we can get a feel for how and why a transmission line antenna works.

Consider a folded dipole made of twin-lead transmission line (**fig. 1**). This type of feedline typically has a 300-ohm characteristic impedance. We can think of this antenna as being driven by our transmitter, an unbalanced RF source voltage. A common and useful technique used to analyze transmission lines is the superposition principle, where the original source voltage is replaced by several different sources which, when combined, add to give the equivalent voltage of the original source. Superposition is used to reconfigure the folded dipole as shown in **fig. 2**. By breaking down the superposition model, it is possible to construct and identify distinctive modes that characterize the behavior of the antenna. **Figure 3A** shows "push-push" or even-mode feeding, in which both wires in the twin-lead transmission line are excited by the same voltage, and have currents traveling in phase. **Figure 3C** illustrates "push-pull," or odd-mode feeding, where the two wires of the twin-lead have currents traveling in opposite directions at any time. The impedances presented by the even and odd modes in terms of the excitation voltage and currents are easily found with the superposition model.

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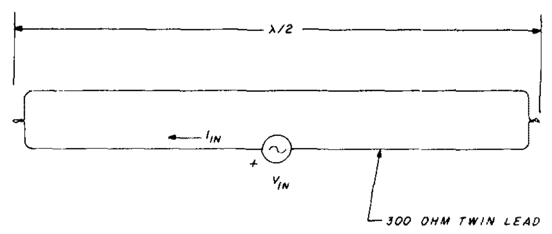

fig. 1. Folded dipole antenna.

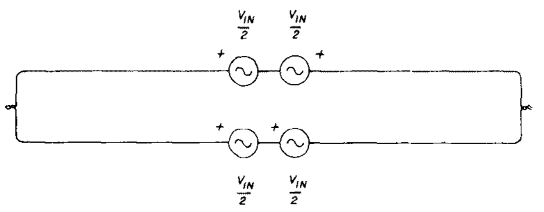

fig. 2. Equivalent folded dipole model using superposition.

For the even mode case:

$$Z_{even} = \frac{\left[\frac{V}{2}\right]}{I_{even}} = \frac{V}{2I_{even}} \; (even\ mode) \quad \textbf{(1)}$$

Since the pair of voltage sources in push-push are similar to just a single source voltage V/2 driving two parallel strands of wire (assuming the twin-lead spacing is much less than a wavelength), the even-mode impedence is that of a "wide" dipole *(Z_{even} = 50 to 75 ohms).* This simplification is shown in **fig. 3B**. Because the even (push-push) mode does the radiating, it is sometimes called the antenna mode. Note that the value of current in each of the transmission line wires is half of the total even-mode current.

For the push-pull case, the odd mode impedance is given by:

$$Z_{odd} = \frac{\left[\frac{V_{in}}{2}\right]}{I_{odd}} = \frac{V_{in}}{2I_{odd}} \; (odd\text{-}mode) \quad \textbf{(2)}$$

Z_{odd} is the parallel combination of the impedances of each of the short-circuited ends of the folded dipole, reflected 1/4 wavelength back to the center feedpoint. Recall that a short-circuited transmission-line offers a near infinite impedance when the source is placed one-quarter wavelength from the short. The odd mode (sometimes known as the transmission line mode) impedance is made very high in this manner. Instead of short circuits, resistors can be placed at various nodes to alter even and odd mode impedances, as well as current distributions. This is sometimes done with rhombic antennas and vee beams. For our folded dipole example, we can observe that the antenna mode offers an impedance to RF on the order of a dipole antenna, whereas the transmission line mode offers extremely high resistances to RF.

Specifically, the input impedance to the antenna is easily computed (using superposition) as:

$$Z_{in} = \left(\frac{V_{in}}{I_{in}}\right) = \frac{V_{in}}{\left[(I_{even}/2) + I_{odd}\right]}$$

$$= \frac{V_{in}}{\left[\left(\frac{V_{in}}{4Z_{even}}\right) + \left(\frac{V_{in}}{2Z_{odd}}\right)\right]} \quad \textbf{(3)}$$

$$= \frac{4Z_{even}Z_{odd}}{(2Z_{even} + Z_{odd})}$$

Observe that for Z_{odd} very large (as is the case here):

$$Z_{in} = 4Z_{even} \cong 300\ ohms$$

We find that not only does this transmission-line antenna radiate, but it also has an input impedance of four times that of a conventional dipole. Conveniently, this structure can be fabricated out of 300-ohm twin-lead, and can also be fed with 300-ohm twin-lead, providing a good match to a 300-ohm receiver. In the case of Yagi antennas, where the mutual impedance effects drop the antenna feedpoint to below 20 ohms, a folded driven element can be used to increase feedpoint resistance by roughly a factor of four.[2]

using the ground as an image

By using the ground to electrically provide half of the antenna system, we can think in terms of a single wire above ground. **Figure 4** shows something that looks like a folded-dipole using an image. Note that the wire height must be much less than a wavelength for the transmission line principles to hold. Since horizontal image currents always travel in opposite directions as do the wire currents, the horizontal portion of this structure, unlike the original folded dipole, will tend to cancel out in the far-field (i.e., the even mode impedance for this antenna is extremely high). The vertical shorting segments, however, will provide a vertical radiation pattern, enabling this antenna to emit and receive RF energy.

Closer inspection reveals that the antenna of **fig. 4A** is identical to the odd-mode excitation of the original folded dipole. Recall that the odd-mode impedance was calculated by considering the parallel combination of the two transmission lines transferred back to the feedpoint. By this technique, it is easy to predict that the input impedance of the structure in **fig. 4A** will be very high, and the radiation will be due primarily to the short vertical segments at the ends of the structure.

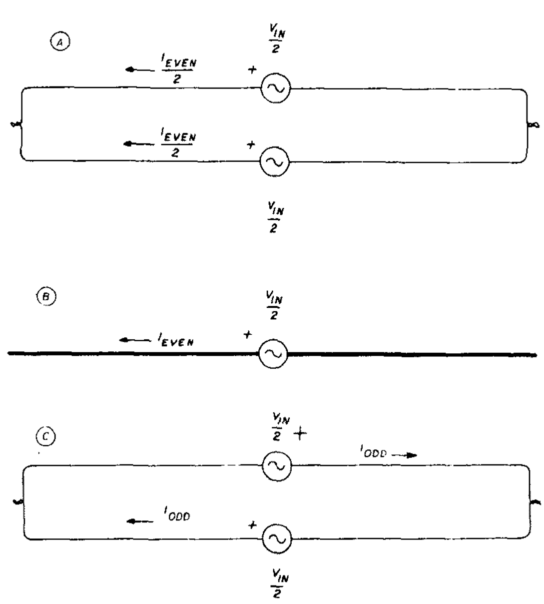

fig. 3. Even and odd antenna mode configurations for a folded dipole: (*A*) even-mode excitation; (*B*) even-mode simplification — dipole antenna; and (*C*) odd-mode excitation.

Now suppose instead of folding both sides of the antenna to ground, we open one of them and move the feedpoint to a "strategic" location **fig. 4B**). This type of antenna is known as a low-profile antenna, and has effectively been used at low frequency (LF) and medium frequency (MF) bands, as well as at microwave frequencies.[3]

low profile antenna

To calculate the input impedance of the low profile antenna at a particular feedpoint we need only deal with the odd mode, since, as was the case for the antenna in **fig. 4A**, the even mode offers an extremely high impedance because of image current cancellation.

Again, we must combine in parallel the impedances of the open and shorted (folded) sides of the structure. For any transmission line, input impedance values may be found by:

$$Z_{short}(x) = Z_0 Tanh(\gamma x) \text{ and } Z_{open}(y) = Z_0 Coth(\gamma y) \qquad (4)$$

where *Tanh*( ) and *Coth*( ) are hyperbolic trigonometric functions, Z_0 is the characteristic impedance of the transmission line, x is the distance from the feedpoint to a short-circuit termination, y is the distance from the feedpoint to an open-circuit termination, and γ is the complex propagation constant of the transmission line, made up of a real attenuation factor, α, and imaginary term, $j\beta$, representing the emission wavelength of the source ($\gamma = \alpha + j\beta$).

The characteristic impedance of the single wire (and its image) is dependent upon many factors. These include height above ground, ground conductivity, and moisture of the air, to cite just a few. At 1.8 MHz, an approximate value for the characteristic impedance of a wire 6 meters above ground is about 800 ohms. The value of Z_0 is really not important, though. The success of this antenna lies in the parameter γ.

Naturally occurring losses in the ground and in the wire cause some slight attenuation in electromagnetic waves as they propagate through the line. This attenuation, α, is expressed in units of relative voltage decrease per unit length (dB/m), and yields the real part of γ. It is instructive to compare a lossy and lossless model of the low-profile antenna to see exactly how it loads.

For lossless transmission lines, where $\gamma = j\beta$, the expressions for short-circuit and open-circuit transmission lines simplify to:

$$Z_{short}(x) = jZ_0 Tan(\beta x) \text{ and}$$
$$Z_{open}(y) = -jZ_0 Cot(\beta y) \qquad (5)$$

where *Tan*( ) and *Cot*( ) are trigonometric functions, and j is the imaginary operator, or a 90-degree phase shift. For this ideal case, the parallel combination of the open and short-circuited line yields an imaginary result for any value of x or y ! Since it is impossible to deliver power to a purely reactive load the SWR is infinity for the ideal case. However, when losses are considered, it is possible to solve for values of x and y which yield a purely real Z_{in}. This indicates that we are using the naturally occurring losses of a transmission line to provide a purely resistive RF load for our transmitter! The end result is an antenna that can be made to resonate at any real impedance, provided the correct lengths of open and short-circuited transmission line are used.

implementation

Solving for the lengths x and y is much too impractical because of the many variables that exist at an antenna site. Trial and error is the easiest way to "zero in" on the particular lengths needed for a desired impedance and a given configuration. For a 50-ohm antenna impedance, I wound up with the dimensions shown in **fig. 5**. Only four tries were required to get the SWR below 1.5:1 in the 1800 kHz to 1850 kHz band, pruning only the longer (open) length of wire. I also discovered that other configurations are possible, at the expense of some bandwidth (**fig. 5B**). Since different locations will use slightly different configurations, it is impossible to derive explicit formulas for the

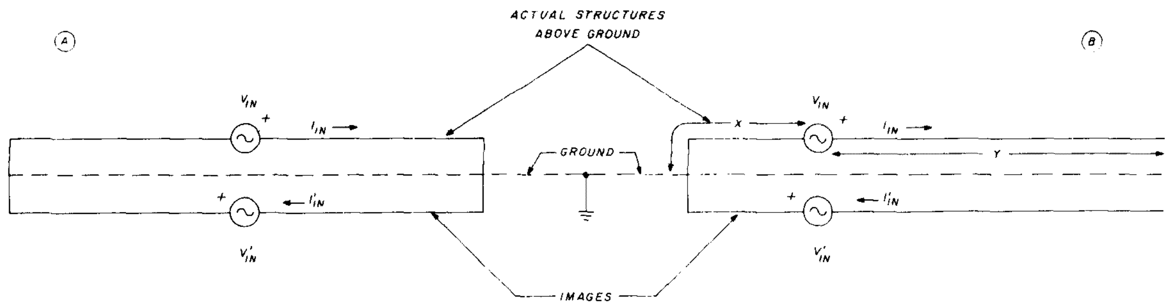

fig. 4. Example of antennas with images: (*A*) pseudo-folded dipole with image; (*B*) low-profile antenna.

wire lengths; however, it is safe to say that the open-circuit length will not exceed 40 meters (0.24 λ), and the short-circuit length should not exceed 9 meters (0.05 λ). To tune the antenna, start with these lengths and trim the longer wire (open-circuited transmission-line) by removing 0.5 meter lengths of wire until the SWR approaches 3:1 at the frequency of interest. Then, very finely prune both the open and short-circuit lines for best SWR. Small lengths of wire may have to be re-inserted after course pruning to optimize the match to the transmitter. It is important to be sure that all antenna SWR measurements are made with the antenna at its operating height, as the wire height above ground critically effects the tuning (in fact, this is another parameter that can be varied in the pruning process to provide the best match). It should be possible to achieve an SWR well below 1.5:1 if patience is exercised in trimming the antenna.

Wire heights from 2 to 8 meters make the transmission line approximation valid, although I would think that heights greater than this could also be made to resonate. The antenna feed system is simply a random length run of RG-58U coaxial cable. The open-circuit wire is connected to the coax center conductor, while the coax braid is soldered to the short-circuit wire, which is terminated at a ground stake (**fig. 6**).

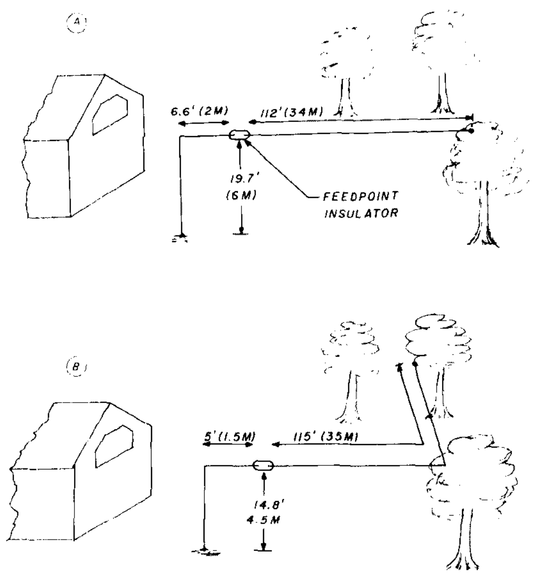

fig. 5. Low profile antenna dimensions for 160-meter operation at N9NB: (*A*) low-profile antenna for 1850 kHz – 100 kHz bandwidth (dimensions: x = 8 meters, y = 34 meters, height = 6 meters); (*B*) low-profile antenna for 1850 kHz – 75 kHz bandwidth (dimensions x = 6 meters, y = 35 meters, height = 4.5 meters).

antenna performance

I was able to obtain an SWR of less than 2.0:1 from 1800 kHz to 1900 kHz using the configuration shown in **fig. 5A** while obtaining the same SWR over a 70 kHz bandwidth for the set-up shown in **fig. 5B**.

My first evening on 160 meters with this antenna was most enjoyable, as I rag-chewed with stations from Delaware to California! Never before have I been able to call CQ and get an answer. It sure beat the RF burns and weak signals I had been used to!

I've worked over 30 states and several DX stations (including two Europeans) in the past month using only a 1-meter long ground pipe and 100 watts of transmitter power. Also incredible is the low receiver noise level. There have been many times when I could copy DX stations, while many other stateside operators could not. This antenna may be of interest to those who *don't* have room for Beverage antennas but want to get away from the received noise characteristics of verticals and dipoles.

The antenna seems to exhibit a moderately high angle of radiation and has a radiation pattern similar to that of a short dipole combined with a short verti-

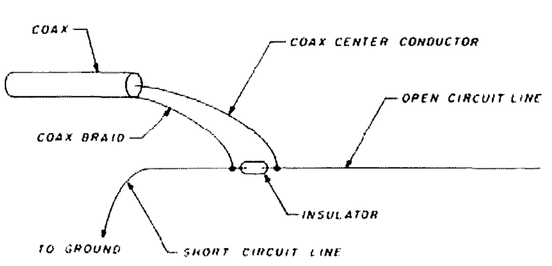

fig. 6. Simple feed method.

cal. The short-circuit line provides a vertical pattern, making this structure similar to a short vertical antenna. The open circuit wire provides some horizontal radiation, and is effective in tuning the antenna to resonance. The efficiency of this antenna is determined primarily by the ground conductivity at the antenna site. Unfortunately, soil is a very imperfect conductor. Ground radials may be used to increase efficiency, although they are not essential. In fact, a poor ground may actually be beneficial as it would prohibit complete cancellation of the horizontal current components in the far-field.

KS9J and WA2JQW have reproduced this structure at their locations using wire heights as low as 6 feet (1.9 meters) and short-circuit wire lengths as short as 8 feet (2.6 meters). They have indicated that low SWR is obtainable using an arbitrary wire configuration, at an arbitrary height, as long as care is taken to prune the antenna patiently.

conclusion

After many frustrating attempts to work the 160-meter band without an adequate antenna, I have finally found something that keeps both me and my transmitter happy. The low SWR allows for operation without an antenna tuner, and the direct coax feedline minimizes RF in the shack. Most gratifying, though, are the many new friends I have made on 160 meters and the enjoyment that contacts on the "gentleman's band" can bring!

Those who are fortunate enough to have plenty of yard space, tall trees, or a tower might not want to use this type of antenna for 160-meter operation. But those of you who think you don't have the room to get on "top band," may want to give this tuner-less, trap-less transmission line antenna a try.

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1. Walter L. Weeks, *Antenna Engineering*, McGraw-Hill, Inc. 1968.
2. Technical Correspondence, *QST*, January, 1984, page 48.
3. Rudolf J. Guertler, "Isotropic Transmission line Antenna and its Toroid-pattern Modification," *IEEE Transactions on Antennas and Propagation*, May, 1977, page 386.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

stacking antennas, part 2

Last month's column discussed the theoretical aspects of stacking antennas.[1] It was pointed out that the optimum stacking distance for two antennas occurs when the beamwidth in the stacking plane is reduced to approximately 50 percent of the original antenna beamwidth and grating lobes are about 13 dB below the main lobe. It was further stressed that the antenna to be stacked should be "clean" (low side-lobe) in order to achieve effective stacking since the 13 dB grating lobe level can't be obtained if the antenna to be stacked has 13 dB or poorer side-lobe levels to start with! Finally, it was shown that *if everything were done correctly*, a gain increase of 2.5 dB, instead of 3 dB, would be a realistic figure every time the number of antennas was doubled.

Several tables and graphs were also presented to enable you to determine the optimum stacking distance for any antenna. While all the electrical parameters are necessary, the practical feeding and physical aspects of stacking antennas are also important. Therefore, this month's column will try to tie the subject of stacking together so that you can choose the optimum configuration for your particular application.

While working on part 2 of this article I noticed that there is one caveat I neglected to point out in part 1. *All the information presented on stacking is based on having no ground reflections* (antennas in theoretical space). However, once an antenna is over 2 to 3 wavelengths above ground (a typical situation on 2 meters and above), the antenna is, for all practical purposes, in free space. At 6 meters there may be a problem since the lower antenna should be at least 40 feet (12 meters) above ground.

stacking configuration

There are literally dozens of ways that antennas can be stacked. However, only a few configurations are typically used by Amateurs.[1] Some of these are shown in **fig. 1**. (For clarity, the mechanical considerations have been omitted from the illustrations.) The simplest stacking configuration places two Yagis in either the horizontal (**fig. 1A**) or vertical (**fig. 1B**) plane. The optimum spacing between the Yagis was discussed in last month's column. I'm often asked the reference point for measuring the spacing: it is the distance between the current points — usually the boom on a Yagi. In the special case of the loop Yagi, it is the center-to-center spacing between the loops.

One of the most popular stacking configurations for higher antenna gain is the "quad" or "box" shown in **fig. 1C**. This is usually a simple mechanical arrangement and has almost identical beamwidth in both the vertical and horizontal planes.

An often-overlooked configuration is an array of six Yagis (**fig. 1D**), which yields a theoretical improvement of 1.75 dB and a typical increase of 1.5 dB over a four-Yagi array. This is only about 1 dB below an eight-Yagi array but with a 33 percent smaller area! This configuration is only recommended using vertical stacking as shown and has a more complex mechanical structure. However, it is particularly recommended for those who can't double their arrays but could expand a four-bay array to six antennas with minimum cost and mechanical impact. Other common high-gain configurations especially popular with EME'ers are shown in **figs. 1E, 1F**, and **1G**. These arrangements are usually the easiest to realize mechanically when very high gains are required.

vertical versus horizontal stacking

Last month's column discussed the problems associated with the side lobes in the H-plane (vertical) of a typical Yagi antenna and showed that they are normally 2 to 3 dB worse than in the E (horizontal) plane. Hence, Yagi antennas must often be stacked closer than desired in the vertical plane to control the vertical grating lobes. Consequently, vertical stacking may yield slightly less gain increase than horizontal stacking.

However, despite these negatives, there are reasons for stacking Yagis vertically. First, there is sometimes only one vertical mast available. Secondly, when very high gain is needed such as for EME, four or more antennas may be required and hence stacking some or all of the antennas vertically is often desirable. *Remember that the array beamwidth decreases only in the plane of the stacking.* Hence, if antennas are stacked verti-

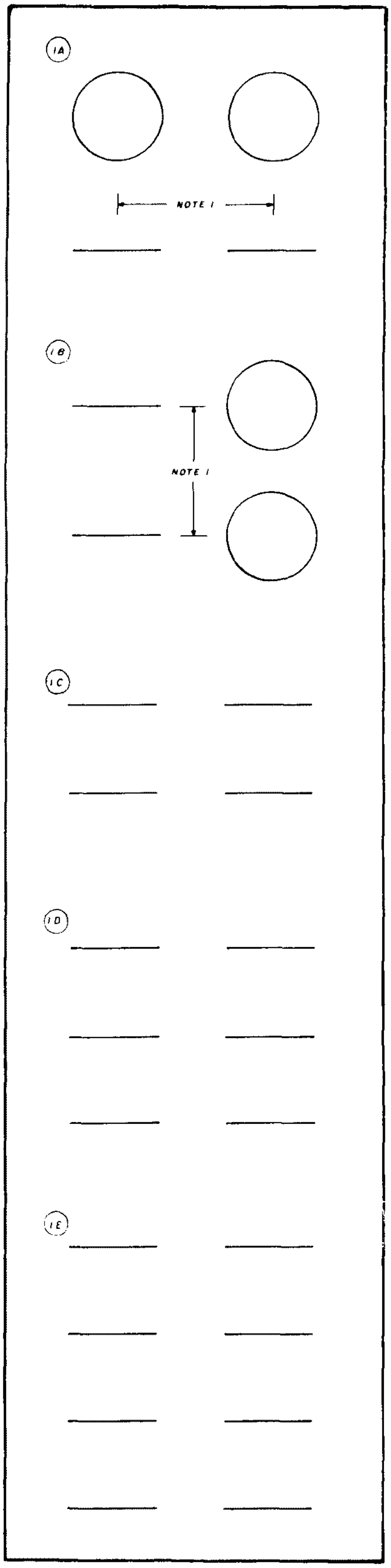
cally, the horizontal beamwidth remains the same. This is particularly desirable when you don't want to "miss" stations that are slightly off the main beam such as when using tropospheric propagation. Likewise, vertical stacking is especially desirable when the signals may be up to 5 or 10 degrees off the great circle path such as in meteor scatter communications. Conversely, horizontal stacking is desirable for auroral propagation since the signal returning from the auroral curtain is usually elevated above the horizon. Horizontal stacking does not affect the vertical beamwidth of the antenna being stacked.

Finally, when high gain is required it is almost impossible to not stack in both planes as shown in **figs. 1C-1G.** If six or eight Yagis are used, you can still tailor the pattern by choosing which plane requires the greater beamwidth.

electrical considerations

After choosing the stacking configuration, there are some important electrical considerations such as choice of feed line and the placement of the power splitter/combiner(s). As mentioned in part 1, losses in the feed harness can severely reduce the gain when stacking antennas. The total loss in antenna gain due to the phasing harness is the sum of the insertion losses from the input of the first power splitter/combiner to the final antenna feed point on the individual antennas (see **fig. 2**). Generally speaking the major losses are in the transmission lines in the phasing harness. Note that the overall antenna gain loss shown in **fig. 2** *will* only be 0.5 dB, *not 2.0 dB* as I sometimes hear! However, while 0.5 dB may sound low, remember that this loss reduces the stacking gain. Furthermore, a 0.5 dB phasing line loss can decrease the receiver signal-to-noise ratio by 2 or more dB when looking at a "cold" sky on 70 cm EME!

From the above discussion, you can see another reason why I get so upset when I see antennas stacked too far



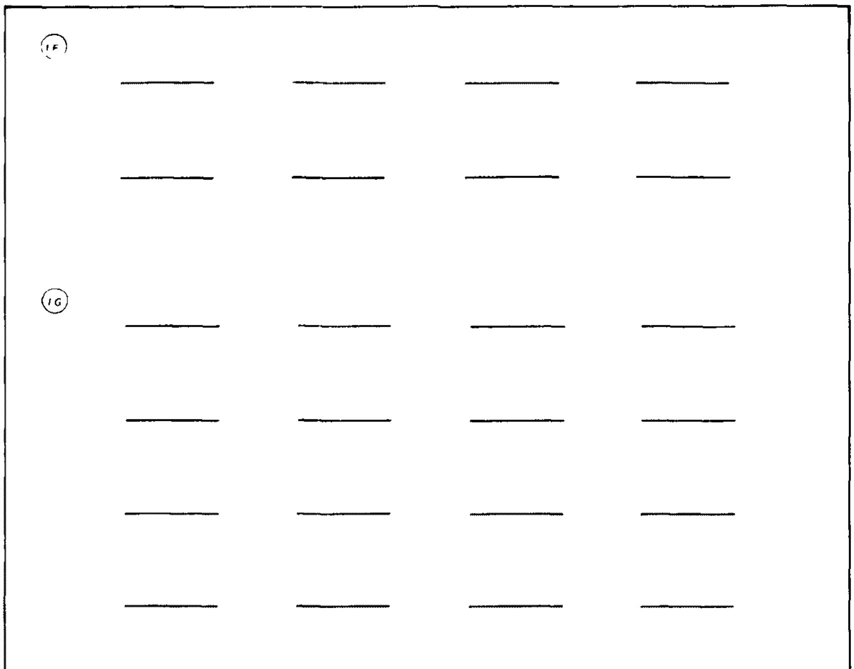


fig. 1. Typical recommended stacking configurations. *A*) horizontal; *B*) vertical; *C*) quad or box; *D*) six-way vertical; *E*) eight-way vertical; *F*) eight-way horizontal; and *G*) sixteen-way symmetrical. Mechanical details have been omitted for clarity. *Note*: reference points for determining stacking distance.

apart. The additional spacing not only increases the grating lobes and decreases the beamwidth unnecessarily but the hoped for gain increase of 0.1-0.2 dB by stacking wider than optimum may easily be offset by the extra phasing line loss. And that isn't all. The clincher against using unnecessarily greater spacing is the increased load that is placed on the rotator and structure.

feeding the array

When stacking antennas, it is common practice to have a separate phasing line on each antenna. This line is then connected to a common splitter/combiner for each 2, 4, or 6 antennas being stacked. Eight-way splitter/combiners are sometimes seen, but I personally feel that they are more difficult to use and can add additional phasing line losses. If coax phasing lines are used, each is *usually run* down the antenna boom to a common power splitter/combiner which is centrally located on the main mast or boom. This method is mechanically sound but often adds excessive feedline losses.

A more recent trend popular with EMEers is to use a "back plane" feed system. In this arrangement each phasing line is routed to the rear of the array rather than down the boom. The power splitter/combiner is then mounted behind the antennas, usually on a separate small boom extending from the main mast. This allows very short low loss phasing lines. The main feedline(s), which are fewer in number, can now be made from a more expensive but low-loss air or foam dielectric coax.

phasing line requirements

For best results all phasing lines should have the same overall electrical length within 22.5 degrees or 1/16 wavelength. This works out to be approximately 1.7 inches (4.3 cm) at 432 MHz. The longer the physical length of the phasing lines, the greater chance of having an unequal electrical length in these lines.

From my personal experience, I can offer the following suggestions. Always make all phasing lines from the same roll or piece of feedline. Stay away from small diameter (e.g., RG-58/U) and low-cost feedlines. If foam dielectric coax is used in a phasing harness, it should be treated carefully; it can "cold flow" when sharp bends are made and is more susceptible to variations with ambient temperature changes.

From practical phasing line measurements I have conducted at 70 cm using a slotted line, I have found that a physical tolerance of 0.25 inch (6mm) on 20 feet (6.5 meters) of 50-ohm RG-213 type coax cable is more than adequate *when using the same spool or length of coaxial feedline.*

I offer another important observation. *Always lay feedlines out in a straight line and measure them in a cool place such as a garage or cellar out of the direct rays of the sun.* In some tests I conducted in my back yard, it was impossible to hold the null on a slotted line constant long enough to take an accurate reading since the sun kept changing the temperature of my coax line and hence varying the length!

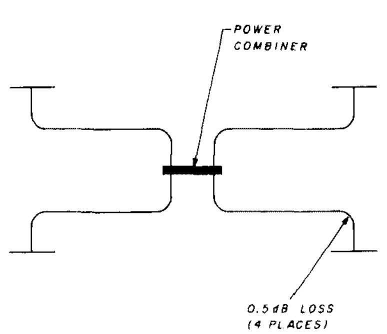


fig. 2. An example of a four-way power split showing the associated line losses. See text for full explanation and overall loss.

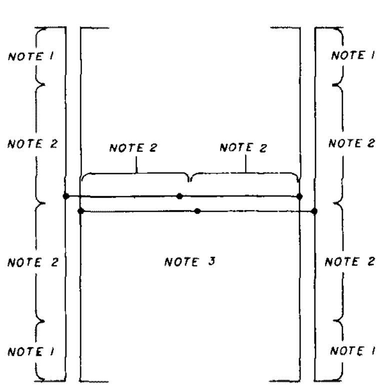


fig. 3. Recommended configuration for an open-wire line type of phasing harness. *Note 1:* quarter-wavelength of 283-ohm open-wire line. *Note 2:* identical lengths of 400-ohm open-wire line. *Note 3:* 200-ohm balanced feed point. Can be connected to a 4:1 balun or connected to another bay using similar matching schemes as shown. See text for further explanations.

open wire lines

I'm frequently asked, "Why not use open wire lines since they are low cost and low loss?" I will not argue with this statement but add the following caveat: *open wire line is a fair weather feedline. Whenever deposits (water, snow, ice, or industrial wastes) build up on the spacers, the impedance and hence the VSWR changes.*

If you decide to use open wire lines, try to keep a low VSWR on all the lines. The extended expanded collinear has a problem in this regard because the VSWR on the feedlines is very high.[2] An array of Yagis with a resistive (non-reactive) impedance is a good open wire candidate. However, use a "*Q*" or quarter-wave matching section at each point where the impedance changes. A recommended example using 200-ohm antennas and feed point is shown in **fig. 3.**

Low impedance (less than 250 ohms) open wire lines are not easy to realize since spacing is close and the conductors have a large diameter. Always use the fewest number of spacers possible and still maintain the mechanical integrity of the line. My experience with extended expanded collinears showed that the maximum power split for open wire lines should be two per junction since four ways can pose symmetry problems and congestion of lines which often leads to unequal splitting and lower gain.

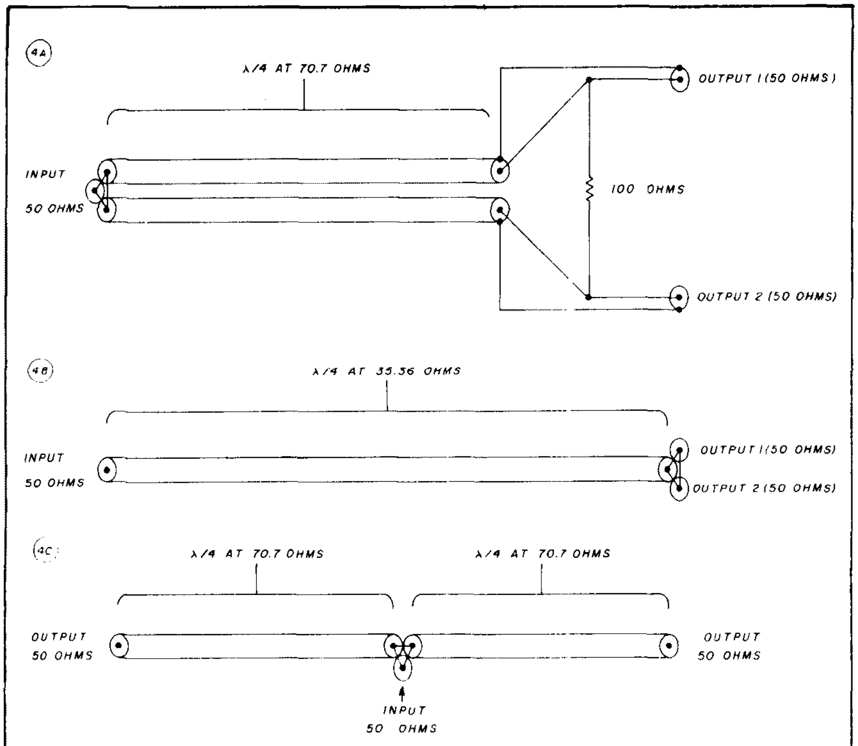


fig. 4. Some typical types of splitter/combiners. ***A*) Wilkinson hybrid; *B*) popular "*Q*" or quarter-wave section; and *C*) half-wave type. (See text for further explanation.)**

phasing line lengths

Finally, I am often asked, "Should the length of the phasing lines be a multiple of half-wavelengths?" This practice apparently started on 2 meters when two antennas were closely spaced and the feedlines were also acting as an impedance transformer.

Ideally speaking, the feedline length is unimportant if the antennas to be stacked have a reasonably low VSWR. However, when antennas are stacked, there is always some mutual impedance. It has been pointed out by Roy Lewallen, W7EL, that whenever any mutual impedance exists between antennas, the ideal phasing line length should be an odd multiple of quarter wavelengths which will enhance the proper distribution of power to each branch of the antenna.[3] He further pointed out that this practice has been in use on TV antennas for many years.[4] Although this may be difficult to realize in practice (especially on 70 cm), *if possible cut your phasing lines to an odd multiple of quarter wavelengths.*

power splitter/combiners

All the foregoing stacking information would be for naught if there weren't a convenient way to do splitting or combining at the antennas in the array. The methods shown in **fig. 3** are fine for open-wire line. Coax, however, requires different handling.

There are two basic types of power splitter/combiner(s) that work well with coax. They are the isolated (hybrid) and the non-isolated quarter wavelength types.

The isolated type shown in **fig. 4A** is often referred to as the Wilkinson power divider.[5] It uses quarter-wave-length lines between a floating resistor. The resistor doesn't consume power unless the loads are unequal or missing. This type of power splitter/combiner is not too popular with Amateurs since the resistors must be able to dissipate high power (at least half the power entering the splitter/combiner) and isolated from ground. Lewallen also indicated that this type of splitter/combiner is not recommended for antenna phasing especially when mutual impedances are present.[6] One of the most popular types of power splitter/combiners used by Amateurs with coax phasing lines is the quarter wave coaxial matching section shown in **fig. 4B**. Its impedance is the geometric mean between the input and output impedances and can be easily calculated using the following equation:

$$Z_{LINE} = \sqrt{Z_{IN} Z_{OUT}} \qquad (1)$$

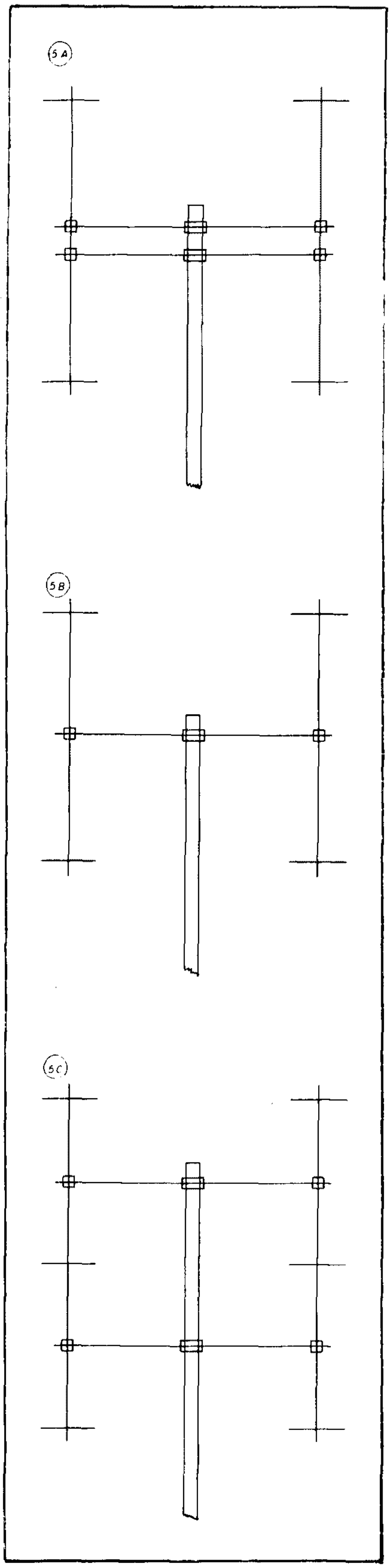
where Z_{IN} is usually 50 ohms and Z_{OUT} is the parallel impedance of the loads. This method can easily be used to split or combine two, three, or four ways. For instance, the impedance of the line should be 25 ohms for a four-way split in a 50-ohm system since the source is 50 ohms and the load is 12.5 ohms (50/4). Likewise a two-way split would require a 35.36 ohm impedance. Both of these impedances are easy to realize using standard square 1 inch (2.54 cm) tubing and hobby shop brass tubing.[7,8,9]

An important attribute of well designed air dielectric splitter/combiners is that they have inherently low loss. There is no reason why they can't be extended internally as described in reference 9. Also, it is often forgotten that most quarter-wave type of power splitters are usable at the third harmonic. Therefore, a good 2-meter power splitter may be also usable at 70 cm and a 70 cm splitter at 23 cm.

Let us not forget to mention the so called half-wave power splitter/combiner. Actually it is still a quarter-wave type since it consists of two quarter-wave sections back to back as shown in **fig. 4C**. It has the added advantage that as a four-way splitter/combiner, the internal impedance is 50 ohms.

mechanical aspects of stacking

In general, try to use symmetrical



stacking configurations because they are less likely to be distorted by sagging resulting from wind or ice loading and are able to return to their intended shape after any external forces or loads are removed. Furthermore, in symmetrical configurations twisting moments are more likely to only cause a moderate decrease in performance if the mechanical structure varies more than 1/16 wavelength or 22.5 electrical degrees (as discussed earlier). Several configurations are shown in **fig. 5**. Stacking frames should be strong mechanically. Excessive weight should be avoided at all cost because it can cause mechanical distortion. The materials used – as well as their diameter and weight – should be carefully examined. The

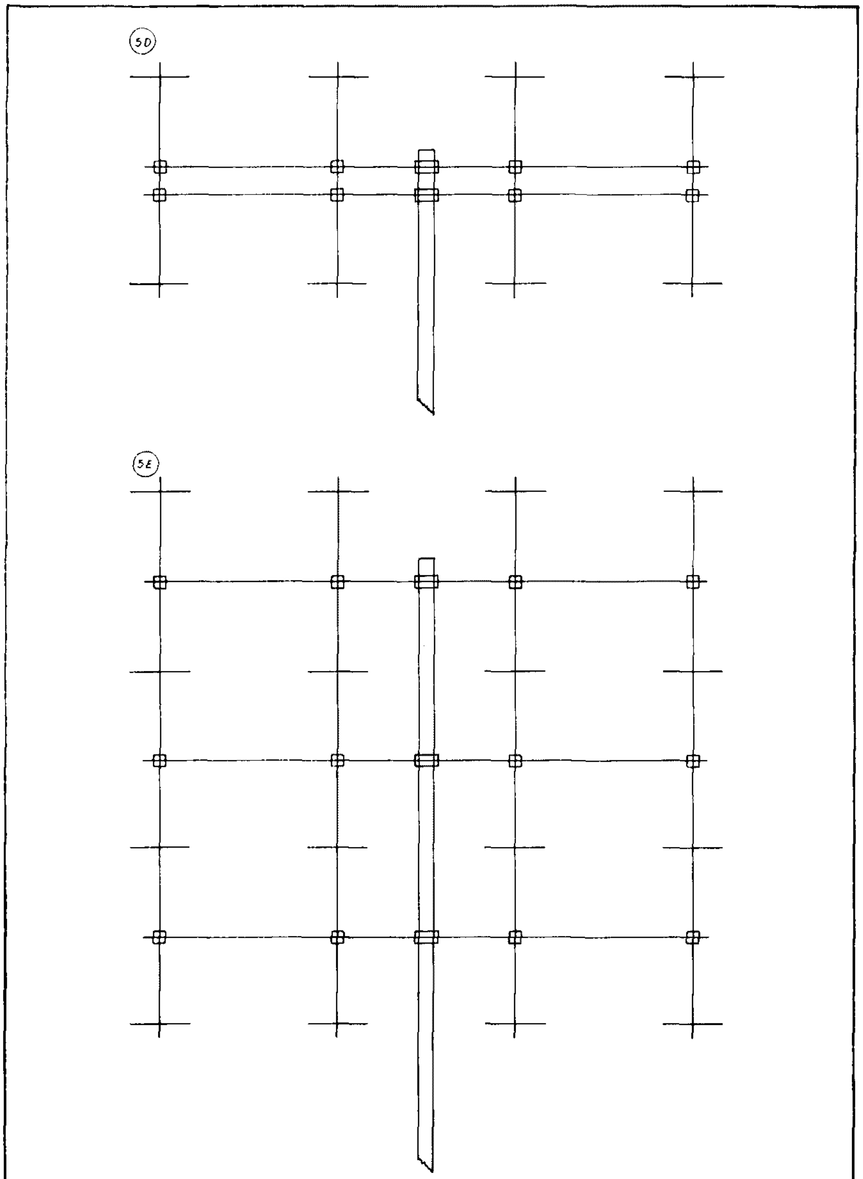


fig. 5. Typical stacking frame configurations for large Yagi arrays. *A*) recommended four-way for a quad of Yagi antennas; *B*) this is not a recommended arrangement since only a single boom connection is used; *C*) recommended six-way stacking frame; *D*) recommended eight-way stacking frame; and *E*) recommended sixteen-way.

table 1. Calculated section moduli for various popular sizes of aluminum tubing using *eq. 2* in text.

outside diameter		wall thickness		
inches	(mm)	inches	(mm)	section modulus
0.250	(6.35)	0.050	(1.27)	0.00134
0.375	(9.53)	0.050	(1.27)	0.00368
0.500	(12.70)	0.062	(1.57)	0.00839
0.750	(19.10)	0.050	(1.27)	0.01810
0.750	(19.10)	0.062	(1.57)	0.02140
1.000	(25.40)	0.062	(1.57)	0.04060
1.000	(25.40)	0.125	(3.18)	0.06710
1.250	(31.80)	0.062	(1.57)	0.06590
1.500	(38.10)	0.062	(1.57)	0.09740
2.000	(50.80)	0.062	(1.57)	0.17870
2.000	(50.80)	0.125	(3.18)	0.32500
2.500	(63.50)	0.062	(1.57)	0.28450
3.000	(76.20)	0.050	(1.27)	0.33610
3.000	(76.20)	0.062	(1.57)	0.41490
4.000	(101.60)	0.050	(1.27)	0.60510
4.000	(101.60)	0.062	(1.57)	0.74930

ARRL *Antenna Book* has an excellent table listing the weight per unit length of most popular aluminum tubing sizes.[10] Furthermore, to prevent bending moments, it is often better to use a slightly thinner wall larger diameter tubing than vice-versa.

The "section modulus" is an excellent way to compare the relative bending moment of various tubing diameters with different wall thicknesses.

$$\text{section modulus} = 0.098\,[(D^4 - d^4)/D] \qquad (2)$$

where D is the outside and d is the inside diameter of the tubing. Typical values of section moduli on some popular tubing sizes are shown in **table 1**. Note, for example, that a 3 inch (7.6 cm) diameter tube with a 0.050 inch (1.27 mm) wall has a higher section modulus and is lighter weight than a 2 inch (5 mm) diameter tubing with a 0.125 inch (3.2 mm) wall thickness. Other values not listed on the table can be easily calculated using **eq.** 2.

Mechanical symmetry of the stacking frame also tends to keep all antenna pattern blockages equal. Likewise, wherever possible, the phasing and feedlines should be neatly and symmetrically dressed and secured to a boom. Stacking frames should have several points of support. Sometimes tower sections are used for the main boom on large arrays! Judicious use of guy wires preferably at right angles to each other with adjustable turnbuckles is also recommended.

Finally, try to pass booms and masts at a mid-point between antennas rather than adjacent to them. If the material or boom used in the stacking frame passes through an antenna pattern, it will have little effect as long as it is at right angles to the plane of polarization and/or is less than 1/10 wavelength in diameter when passing through in the plane of polarization.

other configurations

So far I have been discussing the common arrays. There are many other possibilities. When really high gain is required, a large array of antennas may be constructed. Amateur arrays using 24 Yagis are already in use on 2 meters and 70 cm and a 32-Yagi 2-meter array is under construction! Basically, these arrays are configured using combinations already mentioned. For instance, a 24-bay array can be constructed by stacking three eight-Yagi bays per **fig. 1E** along a main boom. A single three-way splitter/combiner at the center of the array combines the three eight-bay sub-arrays.

Besides the monstrous mechanical problems associated with very large arrays, feedline losses and phasing errors are probably the biggest source of performance degradation. Always try to use the largest and lowest loss phasing lines available. Finally, before constructing a large array, consider whether you are willing to accept all the mechanical risks — it will be large enough! Failure to do so may reduce performance to the point that you may have no more gain than with a smaller array!

final checkout

Whenever stacking is employed, each item should be checked in a methodical fashion. First off, the VSWR of each antenna in the array should be low, preferably 1.2:1 maximum. If coaxial lines are used, *each antenna should be VSWR tested after the antennas are mounted in the array*.

Next, connect the individual antennas into the associated power splitter/combiner in each grouping and test the VSWR again. Then, if applicable, test the entire array.

Be absolutely certain that all antennas are fed on the same side of the array so that 180-degree phase reversals do not occur. Preferably, mount all antennas the same way. If you decide to mount some upside down, make sure to reverse the feed attachment point. Recently, one major antenna manufacturer had a connector plate reversed on some of their antennas.[11] This did not affect performance until two antennas were stacked. Phase reversals of this type can cause radiation patterns to skew or even produce nulls where radiation would normally be present.

Finally, after final assembly and testing, measure antenna azimuthal and elevation patterns, if possible, using the methods described in reference 11. If the antenna pattern does not peak right on boresight (within 1 to 2 degrees) or if the anticipated beamwidth or level of the side lobes is not the value anticipated, as discussed earlier, recheck all electrical and mechanical parameters to find the problem.

summary

In this two-part series, I have tried to cover the major electrical and mechanical considerations required for properly stacking antennas for increased gain. Stacked antennas should be used only when the gain required is beyond that attainable using a single antenna. Remember at the outset that you will be lucky to achieve 2.5 dB of gain increase for every doubling of the array size.

Before building an array that requires stacking, plan carefully. Review both parts of this article several times. Many alternative configurations have been discussed, with pros and cons. Stacking antennas is not a simple job and there are many pitfalls. Both electrical and mechanical decisions must be made. If properly executed, the results can be rewarding. Hopefully the material presented will be useful in building your new super-high-gain array!

acknowledgements

I would like to thank C.J. Beanland, G3BVU, for introducing me to the use of section modulus when comparing different tubing sizes.

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important VHF/UHF coming events

May 1:	*ARRL 70-cm Sprint Contest*
May 3-5:	*West Coast VHF Conference, Sunnyvale, California, (Contact W6RXQ for information)*
May 4:	*EME perigee*
May 4:	*1300 UTC, predicted peak of the Eta Aquarids meteor shower*
May 9:	*ARRL 23-cm Sprint Contest*
May 17-19:	*Eastern VHF/UHF Conference, Nashua, New Hampshire, (Contact W1EJ for information)*
May 19:	*ARRL 6-Meter Spring Contest (tentative date)*
June 1:	*EME perigee*
June 5:	*1930 UTC, predicted peak of Arietids meteor shower*
June 8-9:	*ARRL VHF QSO Party*
June 29-30:	*SMIRK 6-Meter Contest*
June 15:	*0400 UTC, predicted peak of June Lyrids meteor shower*
June 21:	*Mean date of the two-month annual peak of sporadic-E propagation*
June 29:	*EME perigee*

ham radio

RF transmission cable
for microwave applications

Detailed discussion examines all aspects of the vital link between radio and antenna

The correct selection of RF transmission cable requires proper analysis of the electrical and physical parameters of the system.

The amateur microwave enthusiast should be aware that even though materials and dimensions specified by manufacturers are usually accurate, changes in physical and environmental conditions as well as different types of manufacturing equipment or different manufacturing conditions can lead to cable with **substantially** different performance characteristics.

VSWR uniformity

The VSWR of a cable assembly is the summation of reflections due to the connectors, the connector termination technique and the cable. The VSWR of the cable is the summation of random and periodic reflections within the cable, most commonly caused by variations within the processing equipment. The VSWR **will** vary with frequency. A common occurrence is the VSWR "spike" which is illustrated in **fig. 1.**

characteristic impedance

The characteristic impedance of a coaxial cable is determined by the ratio of the inner diameter of the outer conductor to the outer diameter of the inner conductor and by the dielectric constant of the insulating material between the conductors. Select impedance to match your system requirements.

The most common values for coaxial cables are 50, 75, and 95 ohms. Other impedances from 35 to 185 ohms are available in coaxial configurations, but these are normally of interest only to the industry.

Note that the actual input impedance at a particular frequency may be quite different from the characteristic, or surge impedance of the cable due to reflections in the line. The VSWR of a particular length of cable is an indicator of the difference between the actual input impedance of the cable and its average characteristic impedance.

The impedance **will** vary along the length of the cable. Variations of 5 percent are common and some mil spec cables are manufactured to 2 percent tolerance.

capacitance

Capacitance values for standard coax lines depend only on cable impedance and dielectric material.

nominal capacitance pF/foot	cable types
30.8	50 ohm solid polyethylene
25.4	50 ohm foam polyethylene
29.4	50 ohm solid PTFE (Teflon)
20.6	75 ohm solid polyethylene
16.9	75 ohm foam polyethylene
19.5	75 ohm PTFE
16.3	95 ohm solid polyethylene
13.5	95 ohm air space polyethylene RG62B
15.4	95 ohm solid PTFE
10.0	125 ohm air space polyethylene RG63B
6.5	185 ohm air space polyethylene RG114A

capacitance and impedance stability

The capacitance and impedance of long lengths of cable will exhibit very little change over their operating temperature ranges (less than 2 percent). Semi-flexible foam dielectric cables normally exhibit the least change in short cable lengths at frequencies over 1 GHz, although the VSWR can vary significantly if dielectric movement at the connector interface occurs.

average CW power rating

Coaxial cable power ratings must be derated by correction factors for the ambient temperature, altitude (admittedly not of much concern to the Radio Amateur) and VSWR encountered in a particular application. High ambient temperature and high altitude reduce the power rating of a cable by impeding the

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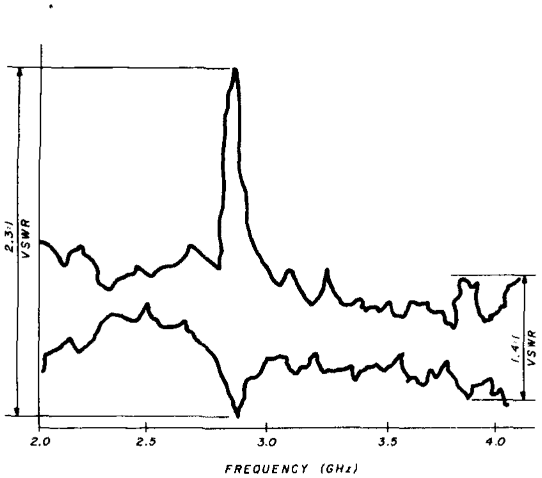


fig. 1. Dual trace pattern illustrates VSWR excursion versus frequency for a 16-foot (5 m) length of 3/4 inch (1.9 cm) 50-ohm cable.

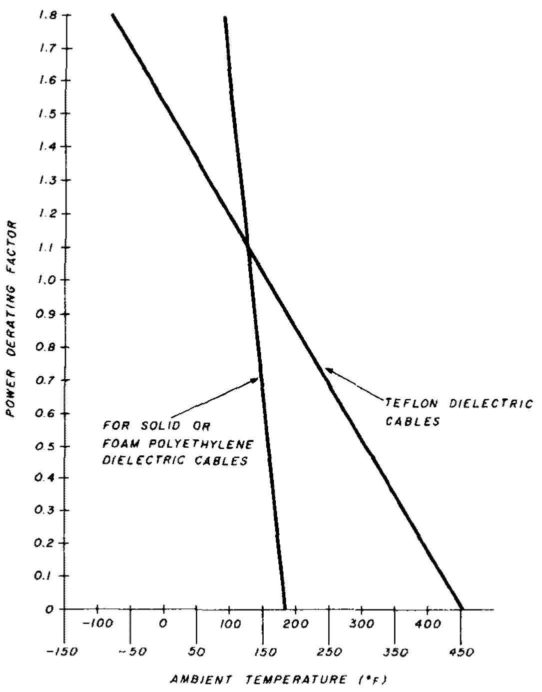


fig. 2. Temperature correction for polyethylene and teflon cables.

heat transfer out of the cable. VSWR reduces power ratings by causing hot spots.

To select the cable construction for a particular requirement, determine the average input power at the highest frequency from your station's requirements.

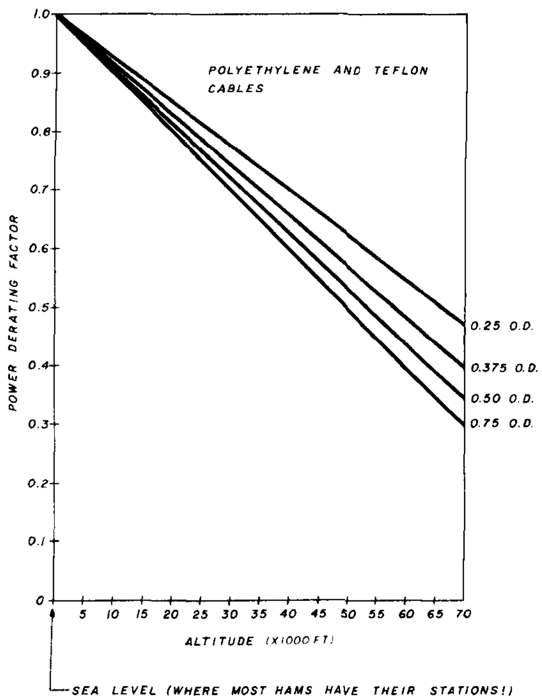


fig. 3. Altitude correction for polyethylene and teflon cables.

Then determine the effective average input power with the following formula:

$$\textit{effective power} = \frac{\textit{average power} \times (\textit{VSWR correction})}{(\textit{Temp. correction}) \times (\textit{Alt correction})} \quad \textbf{(1)}$$

Temperature and altitude corrections are illustrated in **figs. 2** and **3**.

$$\textit{VSWR correction factor} = 1/2(VSWR + 1/VSWR) + 1/2K' \bullet (VSWR - 1/VSWR) \quad \textbf{(2)}$$

(K' is shown in **fig. 4.**)

maximum AC operating voltage

A cable cannot operate continuously with corona because it causes noise generation, dielectric damage and eventual breakdown. The maximum operating voltage must be less than the corona level (extinction voltage) of the cable. This is not to be confused with the dielectric strength of the cable, which is the test voltage applied for one minute during manufacture.

Maximum operating AC (RMS) voltage levels or peak voltages are given for each type of cable in many manufacturer's catalogs. Usually the maximum permissible DC voltage level is 2.5 to 3 times the AC level.

To determine the actual RMS value, divide peak voltage by 1.4. To determine the peak voltage, multiply RMS by 1.4. Then determine the effective input voltage by multiplying the actual input voltage by the square root of VSWR.

The cable you select should have a maximum operating voltage specification greater than the effective RMS voltage.

attenuation

The attenuation of any cable may not change uniformly as the frequency changes. Random and periodic impedance variations give rise to different attenuation responses. Narrow band attenuation spikes can occur.

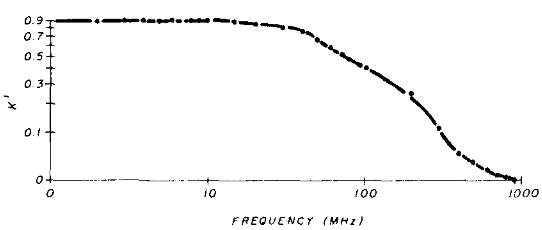


fig. 4. VSWR correction (multiplier K') for average curve.

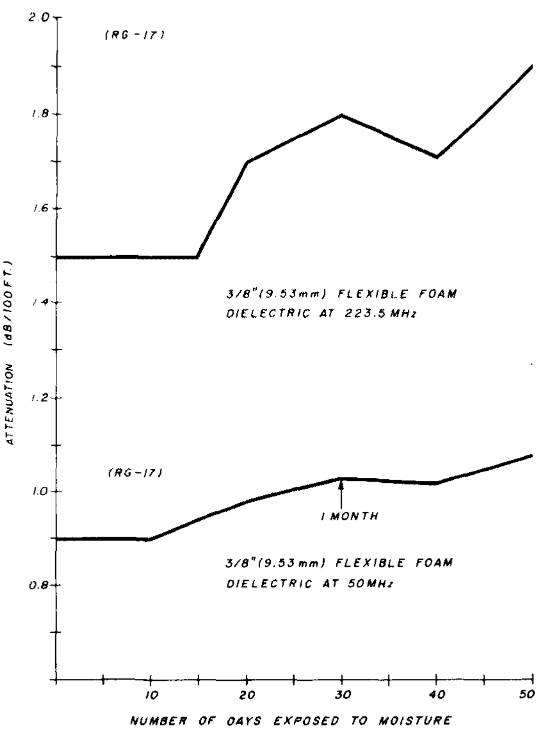


fig. 5. Cable attenuation increases with both frequency and duration of exposure to moisture.

The attenuation of braided cables can increase with time and flexure. The change with time can be caused by corrosion of the braided shield, by contamination of the primary insulation caused by chemicals in the cable jacket, and by moisture penetration through the jacket. Attenuation degradation is more pronounced at frequencies above 1 GHz. Cables having bare copper and tinned copper braids exhibit far greater attenuation degradation than do cables having silver plated copper braids. Refer to **figs. 5, 6,** and **7.**

The following "rules of thumb" apply in Amateur service:

Tin-plated braids. Below 1 GHz, cables manufactured with tin-plated braids have at least 20 percent more attenuation than copper braids in the "as manufactured" condition, but are more stable than bare copper-braided cables.

Foam polyethylene. Flexible braided cables with foam polyethylene dielectrics have approximately 15 percent less attenuation that solid polyethylene cables of the same core size and impedance. However, as many of us have discovered, the attenuation of foam cables will increase if moisture is absorbed. In high humidity environments I suggest that foam cables not be used above 148 MHz. All of these problems can be eliminated by the use of semi-flexible cables; semi-

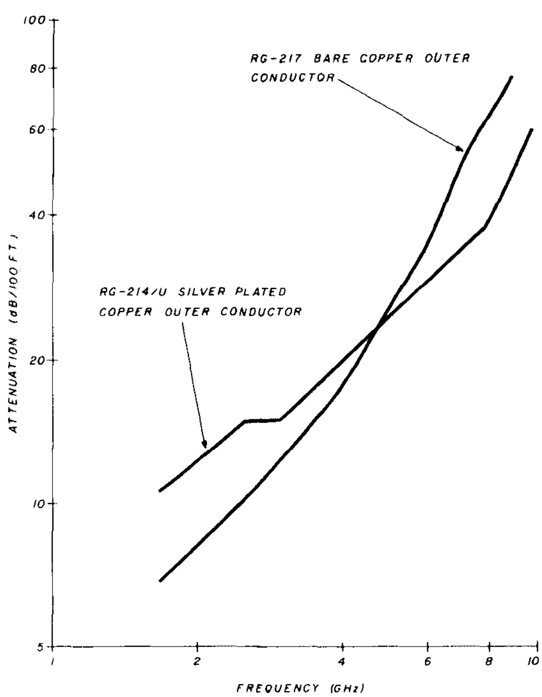


fig. 6. Both RG214 and RG217 exhibit similar attenuation characteristics vs. frequency.

table 1. Formulas common to all coax cable.

$$Capacitance\ (C) = \frac{7.36E}{LOG(D/d)}\ Picofarads/ft$$

$$Inductance\ (L) = 0.140\ LOG(D/d)\ Microhenries/ft$$

$$Impedance\ (Z_0) = \sqrt{\frac{L}{C}} = \frac{138}{\sqrt{E}}\ LOG\ (D/d)\ ohms$$

Velocity of propagation as percentage of speed of light $= \frac{100}{\sqrt{E}}$

$$Time\ delay = 1.016\sqrt{E}\ nanoseconds/ft$$

$$Cutoff\ frequency = \frac{7.50}{\sqrt{E}\ (D + d)} = F_{co}\ (GHz)$$

$$Reflection\ coefficient = \Gamma$$

$$= \frac{Z_r - Z_0}{Z_r + Z_0} = \frac{VSWR - 1}{VSWR + 1}$$

$$VSWR = \frac{1 + \Gamma}{1 - \Gamma}$$

$$Peak\ voltage = \frac{1.15\ S \cdot d\ (LOG\ D/d)}{K}$$

$$\alpha = \frac{0.435}{Z_0\ (D)}\left[\frac{D}{d} \cdot K_1 + K_2\right]\sqrt{F} + 2.78\sqrt{E}\ (P.F.)\ (F)$$

where: α = attenuation in dB/100 ft
d = the outside diameter of inner conductor in inches
D = the inside diameter of outer conductor in inches
S = the maximum voltage gradient of the cable insulation in volts per mil (thousandth of inch)
E = the dielectric constant of the insulation of the cable
LOG = logarithm to base 10
K = safety factor
K_1 = strand factor
K_2 = braid factor
F = frequency in MHz
$P.F.$ = power factor

property of insulating material (dielectric constant "E")

material	dielectric constant "E"
TFE	2.1
polyethylene	2.3
cellular polyethylene	1.4-2.1
polyvinylchloride	3.00-8.00
silicone rubber	2.08-3.50
ethylene propylene	2.24

flexible cable is generally not available to the average ham. Its price is also prohibitive, rising into the *dollars-per-foot* price range.

It is possible however, to use foam polyethylene cable up to 12 GHz. The only catch is that it must be protected from the environment by either running it through a conduit with forced dry air pumped under pressure or sheathing it in a seamless metallic tube.

velocity of propagation

The velocity of propagation of cable is determined primarily by the dielectric constant of the insulating

materials between the conductors. This property is expressed as a percentage of the velocity of light in free space.

cable dielectric	time delay nanosec/ft	velocity percentage
solid polyethylene	1.54	65.9
foam polyethylene	1.27	80.0
foam polystyrene	1.12	91.0
air space polyethylene	1.17	86.0
solid teflon	1.46	69.4
air space teflon	1.16	87.5

cable noise

An area often overlooked by Amateurs is self-generated cable noise, a phenomenon noted whenever a cable is flexed. Both acoustical and electrical noise are generated. This problem can be minimized by properly securing cable to rigid physical structures during antenna system installation.

Most prevalent in the "RG" series cables (RG-8, RG-213, RG-58, etc.) and should be carefully considered by hams who are concerned with feedline performance.

environmental resistance

The life of a coaxial cable depends on many factors other than the materials used in manufacture. The following factors all contribute to cable failure.

UV exposure. Polyethylene-jacketed cable has twice the life expectancy when exposed to direct sunlight as cable manufactured with PVC (Polyvinyl chloride) jackets.

Humidity. All cables experience vapor transmission through their plastic jackets. In Amateur applications it is advisable to install cable where it will not lie in or pass through an area where standing or running water is present.

Salt-water immersion. The electrical characteristics of cable will be rapidly affected if the conductors are exposed to salt water. If you live near a large body of salt water inspect your coax regularly for salt build-up. If your cable should become immersed, try to test the cable with a TDR (time-domain reflectometer) or find someone who can "sweep it" over the frequency range that your antenna system operates.

Underground burial and galvanic action. If you are going to install antenna cable underground, use armored/waterproof coaxial cable or regular "RG" cable installed in conduit or similar protective pipe or tubing.

selection guide

For the Amateur interested in communications through 50-MHz, RG-58 and RG-59 are prudent and inexpensive choices, though their power handling

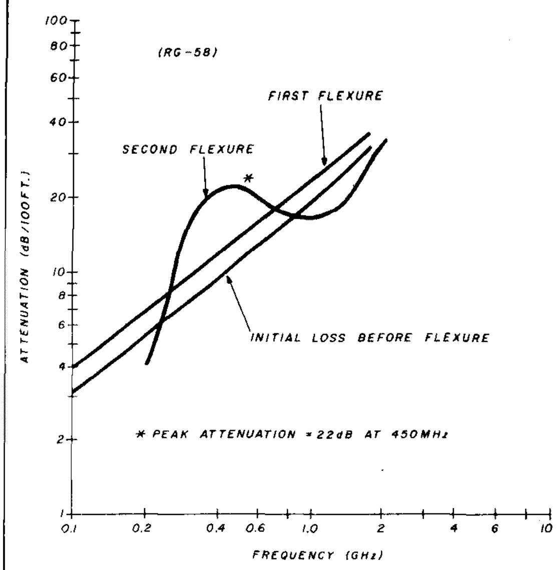


fig. 7. Repetitive flexure of RG-58 cable increases its attenuation characteristics.

capability is limited to 250 watts for RG-58 and 450 watts for RG-59.

A station designed to operate through 148 MHz should incorporate RG-8 or RG-213 coaxial cable. Unfortunately, the 3-dB roll-off point is just below 200 MHz. 220-MHz operation is possible, but 50 percent of transmitted power will be lost in a typical 100 foot run.

Many exotic transmission lines are available for 220 MHz and above. Flexible foam dielectric cable (FM-8 or FM-11, similar to RG-8 and RG-11) is useful up to 450 MHz but very susceptible to moisture penetration.

Flexible low loss cables are manufactured with braids of flat strips of silverplated copper. This helps to lower the VSWR and reduce attenuation above 1 GHz. Type SF-226, for example, is good through 10 GHz and its outside diameter is not much larger than that of RG-8.

For operation above 10 GHz I suggest using corrugated tubular aluminum with foam teflon dielectric, which is available in 3/8-inch OD for operation through 15 GHz. This cable is similar in appearance to the "BX" type of electrical conduit. The bend radius of this cable is 2.0 inches — an important fact to remember during installation.

The most popular coaxial cable among microwave enthusiasts is "hardline" or semi-flexible foam dielectric cable. It is available in various ODs, but for coverage through 22 GHz 1/4-inch OD is your best choice. Although this cable is quite expensive, CATV companies will occasionaly discard it because of minor imperfections or oxidation on the outer jacket. This

discarded cable is suitable for Amateur service. The only real problem with its use and installation is the proper attachment of connectors; this task should be farmed out to your local Amateur microwave expert or to your local cable connector technician!

For use above 22 GHz, tubular copper telfon dielectric cable has *no* radiation loss except at the connector interface. An example is CL-50087T, which is good through 60 GHz. Difficult to work with and very expensive, this type of cable is used in military avionics.

conclusion

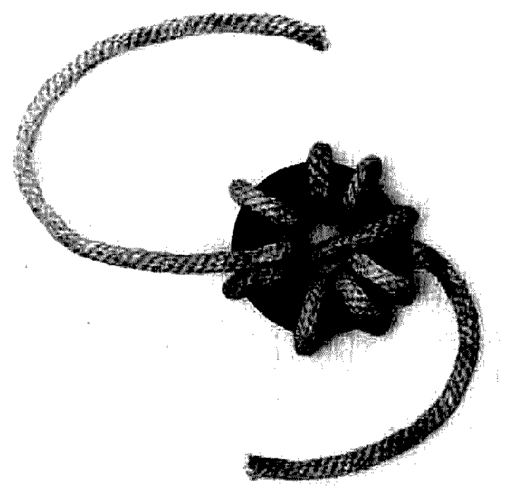
Considering the many electrical and physical characteristics that must be analyzed, selection of RF coaxial cable is really quite complicated. Knowing these parameters is important when attempting to design a communications system for optimum performance.

I am willing to provide additional information on cable, connectors, and installation procedures. Please send requests to my home address, which appears at the beginning of this article. (Please be sure to enclose a legal-size SASE with 3 first-class stamps or IRCs; only inquiries with SASEs can be answered.)

ham radio

balun chop suey

A collection of odds and ends on 1:1 transmission-line baluns



A toroidal core with a super-toroid winding. This core is wound with clothesline for photographic purposes.

In San Francisco's famous Chinatown, I learned a secret: the words "chop suey" mean "odds and ends," which is apparently what that familiar Chinese dish is composed of. This article, a collection of odds and ends about 1:1 transmission-line baluns, is based on comments and questions I have been asked as a result of my previous balun articles. None of the individual comments is long enough to warrant *separate* articles — hence the peculiar title. These odds and ends also refer to balun articles presented by other writers. I hope this discussion will interest balun users and designers alike.

introduction

First, to establish a frame of reference, I plan to discuss transmission-line baluns with input-output impedance ratios of 1:1. The transmission line is wound into an inductance, usually on a ferrite form, either a rod or a toroid.

A balun serves two purposes: first, to provide equal and opposite voltages to a balanced load, and second, to provide isolation between the coax outer conductor and the half of the balanced antenna connected to the outer conductor. Because the *second* is the more difficult problem, I will limit my discussion to it.

The isolation requirement of a balun has been well defined qualitatively by Walter Maxwell, W2DU.[1] Maxwell's article was the first to describe, in the Amateur literature, the concept of *separate currents* flowing on the inside and outside of the outer conductor, representing the signal current and unbalanced currents, respectively. I have found this concept helpful in describing the isolation function of a balun, both in a previous article[2] and in this one. I will expand on Maxwell's comments by giving some numbers to specify isolation quantitatively.

isolation

The isolation function of a balun is to provide a high impedance between the outside of the outer conductor, and *that* half of the dipole *antenna* connected to the outer conductor, without affecting the current flowing on the inside of the outer conductor to its half of the dipole. Very little appears to have been published, at least in the Amateur literature, either quantitatively or qualitatively, on the isolation property of a balun.

Briefly, isolation is necessary to ensure that the signal current flowing on the inside of the outer conductor flows into the antenna, not back down the outside of the outer conductor to ground. Lack of sufficient isolation is one reason antenna currents become unbalanced.

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In its simplest terms, the situation can be considered as a circuit problem. To ensure that the transmission-line current flows into the antenna and not back down the outside of the outer conductor, the impedance provided by the path down the outside of the outer conductor should be many times that of one-half the dipole antenna. For example, if the impedance of the dipole antenna is assumed to be 70 ohms, the impedance of one-half the dipole will be 35 ohms. By making the impedance of the balun to these "outside" currents 10 times this impedance, or 350 ohms in this case, at least 90 percent of the transmission-line current will flow into the antenna and less than 10 percent back down the outside of the coax.

A balun impedance of five times the load impedance (as used above) is an arbitrary figure, although I feel it should be adequate for a general-purpose balun. For a precision balun, one might prefer an isolation of, say, 10 or more times the impedance presented by one-half the total load. In a transmission-line balun, the isolation is provided by the inductive reactance of the winding. The isolation actually provided by a given balun can be determined by measuring the inductive reactance as described in reference 2.

Assuming a linear inductor, as long as the operating frequency is well below the self-resonant frequency of the inductor, the inductive reactance will be directly proportional to frequency. Therefore, there will be a frequency below which the inductive reactance will not be sufficient to provide the required isolation between the antenna and the coax feedline. This determines the low frequency limit of the balun.

The maximum usable frequency of a balun is limited by the stray capacitance across the winding. As the operating frequency increases, a frequency will be reached where the stray capacitance across the balun resonates with the inductance of the winding so that the winding is in parallel resonance. At this frequency the impedance of the balun, and hence its isolation, is the maximum it will be, so that this is a desirable frequency at which to operate. As the operating frequency is further increased, the reactance of the balun becomes capacitive and decreases with frequency until the balun becomes series-resonant. At this frequency the balun impedance is very low and the balun provides virtually no isolation at all. The upper useful frequency limit of the balun is usually between the parallel and series resonant frequencies. The actual frequencies are a function of the construction techniques, the object being to maximize the inductance and minimize the stray capacitance across the winding. (This problem is a subject in itself.)

Fig. 1 shows the inductive reactance versus frequency of a typical transmission-line balun. From a graph such as this, the useful frequency range of the balun can be easily determined for any value of load impedance. The assumed isolation impedance of five times the load impedance, based on VSWR considerations, has been arrived at by others.

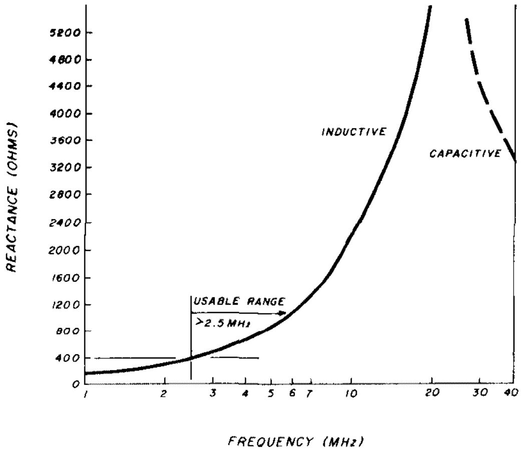

fig. 1. A typical balun reactance versus frequency characteristic. Isolation is at least five times the characteristic impedance from 2.5 MHz to above 40 MHz. (From reference 2.)

To put the isolation property of a balun on a sound technical basis from the user's point of view, I believe balun manufacturers should specify the minimum ratio of the balun winding impedance to the characteristic impedance over the specified frequency range.

testing core saturation

The methods used for testing a balun for core saturation are simple; one method is to wrap a few turns of insulated wire around the balun, as shown in **fig. 2**, and then connect this winding to an oscilloscope to observe the waveform. Gradually increase the output power of the transmitter until a distorted waveform appears. Then back off the power until the waveform again becomes sinusoidal. This represents the maximum peak power a given balun can handle.

A more sensitive version of this method involves connecting the test winding to a spectrum analyzer, instead of an oscilloscope, and measuring the amplitude of the various harmonics. Care must be taken here, however, to make certain the harmonics measured are generated by the balun and not by the transmitter.

An even more sensitive test is to use the RF equivalent of the two-tone test used to measure intermodulation distortion on high-fidelity audio equipment or receiver front-ends. Here two transmitters operating on slightly different frequencies are connected to the balun (and a dummy load) through a diplexer. This test

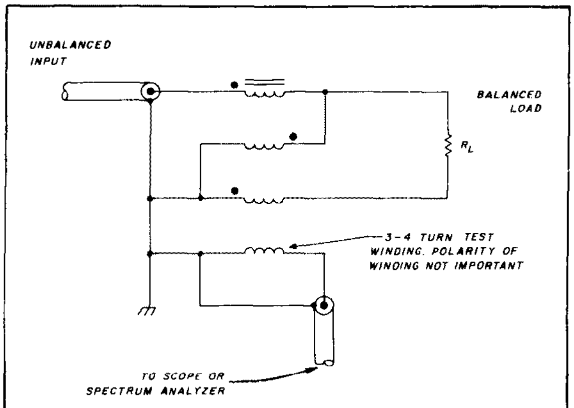

fig. 2. Circuit for core saturation test.

was proposed by Rich Rosen, K2RR, editor-in-chief of *ham radio*; I have not tried it myself.

If saturation is present, it can be corrected by increasing the number of turns, or by using a larger core (cross section), or both.

Increasing the number of turns will also have the effect of improving the performance of the balun at the low end of the frequency range. It will also lower the high-frequency end of the range by reducing the self-resonant frequency of the winding.

If the existing winding efficiently uses the available winding area, it will be necessary to increase the size of the core to accommodate the additional turns unless it is possible to use smaller wire. However, in a transmission-line balun the wire size may be dictated by the desired (or required) characteristic impedance of the winding; the wire diameter and spacing — being parameters — determine the characteristic impedance of the winding.

While the equipment needed and the procedures are relatively simple, the practical problems of testing baluns can be substantial. First, it takes a high-power transmitter and a high-power, preferably balanced, dummy load. Most Amateurs, however, will probably use an unbalanced load because this type is the most readily available. Care must be taken that the load case is well insulated from ground because the case will be at one-half the unbalanced line voltage when used as a balanced load.

The transmitter must be capable of providing the highest power level at which the balun is expected to operate. The balanced dummy load, of course, must be capable of dissipating this amount of power.

A more convenient load, at least if you are testing the balun for your own use, is to use the regular station antenna. (Choose a time when the band is dead!) The problem here is that the balun will be high in the air, requiring long test leads in a high-power antenna field, because RF can be picked up by the line and result in inaccurrate measurements.

transformers or inductors?

One of the principal sources of confusion in balun design, construction, and use seems to be the belief that all baluns are transformers. One well-respected writer recently stated "It is important to recognize that a 4:1 or a 1:1 balun . . . is essentially a broadband transformer." If he had stopped to think about it, I'm sure this writer would have known better. Strictly speaking, any device or collection of components which transforms a balanced line to an unbalanced line *can* be called a transformer; in electronics the expression "transformer" is usually reserved for one particular type of device where all, or part, of the energy passes from input to output by means of magnetic induction. The usual two-winding or three-winding 1:1 transmission-line balun, however, is not a transformer because none of the energy is transmitted from input to output by magnetic induction. This type of balun should not be designed, tested, or used as a transformer. A 1:1 transmission-line balun is an inductor wound with a transmission line and must be designed, tested, and used as an inductor wound with a transmission line. Failure to recognize this difference in design or application is almost certain to lead to disappointment in the use.

Further complicating the situation is the fact that a transmission-line balun can be constructed with impedance transformation ratios other than 1:1, although I have not seen this done in Amateur applications. The common 4:1 balun is really an auto-transformer and not a transmission line device.

A broadband transformer, on the other hand, must be designed in accordance with well-known transformer equations relating the number of turns, the peak voltage, and the allowable maximum magnetic flux density in the core.

A second means of distinguishing between transformer and transmission-line baluns is that with a transformer balun, the relationship between the load impedance and input impedance is the turns ratio squared. With a transmission-line balun, the load-to-input impedance ratio is calculated using the more complex transmission-line equation:

$$Z_{in} = \frac{Z_{ch}\,(Z_r \cos\theta + jZ_{ch} \sin\theta)}{(Z_{ch} \cos\theta + jZ_r \sin\theta)} \qquad (1)$$

There are two problems in designing transmission-line baluns. The principal problem is designing a wideband inductor, i.e., an inductor whose reactance is above a given value — usually five to ten times the load impedance — over the desired range of frequencies. The second problem is designing a suitable trans-

mission line, with the required characteristic impedance, capable of being wound on a ferrite core.

Briefly summarizing, because of the similarity in the appearance and circuit configuration of different types of baluns and transformers, both of which use magnetic cores, there is considerable confusion about their operating characteristics. This in turn leads to improper design and application of the devices. Before criticizing a device, one should make certain it has been properly designed and used.

the super-toroid

The super-toroid balun design was introduced to the Amateur community by Reisert in his article on the two-winding transmission-line balun.[3] The concept of a super-toroid, however, was developed by Tom Gross in the early 1960's; Gross is a designer of precision magnetic components who specializes in devices operating at supersonic frequencies — 20 kHz to 500 kHz and up. The purpose of the super-toroid design is to reduce the sensitivity of the conventional toroidal coil to external magnetic fields.

Contrary to popular belief, a single-layer toroidal inductor is sensitive to external magnetic fields because of what is known as the one-turn effect. With a conventional, single-layer toroid, the winding is spirally wound around the core, as seen in **fig. 3**. Each turn is composed of two winding components: one component is aligned with the center of the toroid. The other component is at a right angle to the radial component and is known as the circumferential component. These two components are emphasized in **fig. 4**. The circumferential component provides the progression of the winding around the core. If there were no circumferential component, each turn would lie on top of the preceding turn.

The circumferential components constitute a single-turn loop antenna which, as is well known, has a response to extraneous signals originating in the plane of the toroid, as shown in **fig. 5**. This is known as the "one turn" effect of a toroidal winding. If the winding has more than one layer, the single-turn effect becomes an N-turns effect, where N is the number of layers.

The super-toroid design eliminates the one-turn effect by winding half the turns in one direction, then running the winding across the diameter of the core to the opposite side and winding the other half of the core in the opposite direction. The voltages induced in the circumferential components are therefore equal and opposite and therefore cancel. The photo shows a core wound in this manner. The winding in this case is clothesline rope to show up better in the photograph.

This technique assumes that the *external* magnetic field is uniform through the diameter of the core and

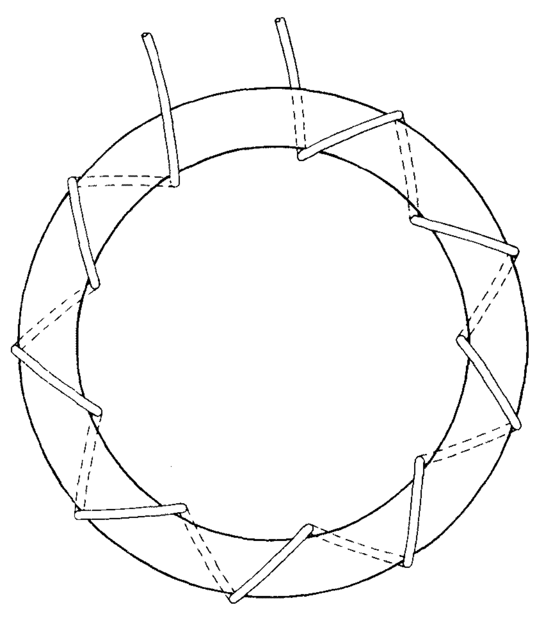

fig. 3. A conventionally wound balun emphasizing the winding pitch, which causes the 1-turn effect.

the core material is homogeneous. If both these conditions are not met, even more complex types of windings must be used to provide isolation from external magnetic fields.

A secondary advantage of the super-toroid design, and one that is much more important to balun designers, is the fact that the effective capacitance across the winding is reduced, which increases the usable bandwidth of the device.

Any inductor consists of stray capacitance as well as inductance. In the design of an inductance, the object, of course, is to maximize the inductance and minimize the capacitance. As seen in **fig. 6**, this capacitance consists of an infinitely large number of incremental capacitances such as the capacitance between adjacent turns, capacitance between turns that are not adjacent, and between the ends of coil. From an engineering standpoint, it is not practical to evaluate an infinite number of small capacitances. Therefore, the "effective capacitance" of an inductor is defined as that of a physical capacitor connected across the ends of the inductor, which stores the same amount of energy as that stored by the incremental capacitors.

The energy stored in a capacitor is proportional to the product of the physical capacitance and the voltage across the capacitor squared: $Energy = \frac{1}{2} CV^2$. If the two plates of a capacitor are at the same potential, the energy stored by the capacitor is zero

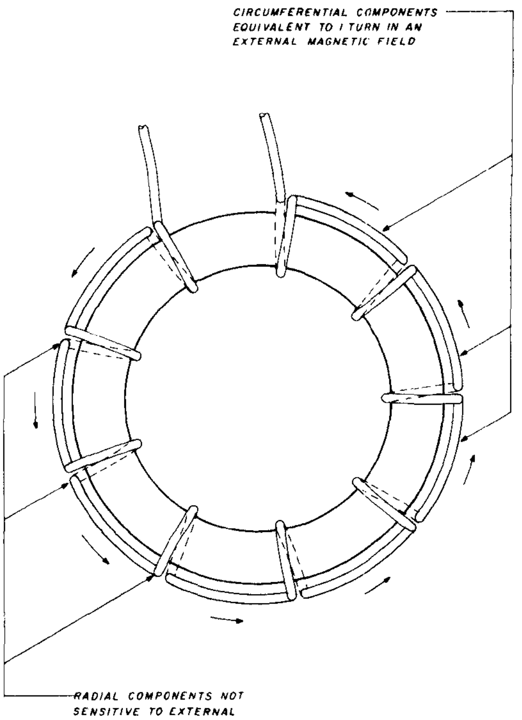

fig. 4. A conventionally wound toroid equivalent to *fig. 3* but emphasizing the 1-turn effect.

and the effectiveness of the capacitor is nil. It is, therefore, important that the portions of an inductance that are at the greatest potential difference have the lowest possible physical capacitance. The ends of the winding are at the greatest potential difference, hence they should have the greatest separation to minimize the physical capacitance.

A super-toroid has the ends of the windings at opposite ends of the core diameter, which is about as far apart as it is possible to place them. A conventionally wound balun, on the other hand, has the ends of the windings adjacent to each other, where they have the highest capacitance. It is, therefore, easy to see why the super-toroid winding would have a lower effective capacitance and hence a greater bandwidth. Then why not use the super-toroid winding exclusively?

For a two-winding balun, the super-toroid winding is preferable. For a three-winding balun, the situation is complicated by the tertiary interconnections between the opposite ends of the two main windings. With a continuously wound toroid, these interconnections can be short and can have a low impedance. With a super-toroid winding, the two leads of the tertiary winding come out on opposite sides of the core, so that both tertiary leads must cross the core. This greatly increases the leakage reactance between the main winding and the tertiary winding and will prove detrimental to proper balun operation above 20 MHz.

Thus, even though the super-toroid winding inherently gives greater bandwidth, it should not be used for a three-winding balun. For three-winding baluns, it is necessary to use a continuous winding to ensure short, low impedance interconnections between the tertiary and main windings.

wide bandwidth?

Numerous *ham radio* readers have questioned the desirability, or necessity, of the wide bandwidths advertised for commercial baluns — typically 3.5 to 30 MHz, and more recently, 1.5 to 56 MHz. Are bandwidths this wide really necessary? Or desirable? The argument is that since few, if any, Amateurs use the same antenna for 80 through 10 meters, much less 160 through 6 meters, why insist that a balun cover a greater, wider bandwidth than the antenna?

My own opinion is that these wide bandwidths are unnecessay and may not be desirable. The antenna installation for a typical Amateur active in the HF region might be a dipole on 80/75 meters or 40 meters plus a triband beam for 20, 15, and 10 meters. If space permits, maybe a 6-meter beam, too. Therefore, a balun that gave high performance, say, from 3.5 to 10 MHz and a second balun optimized for 14 to 30 MHz could be designed to provide higher performance at the extreme edges of these bands than a single wideband balun covering the entire frequency range.

This has been substantiated by my own experience in building and measuring baluns. Wideband baluns — 80 through 10 meters — gave marginal but acceptable performance at the low end of 80 meters and at the high end of 10 meters. The isolation impedance was just barely equal to five times the characteristic impedance of the transmission line.

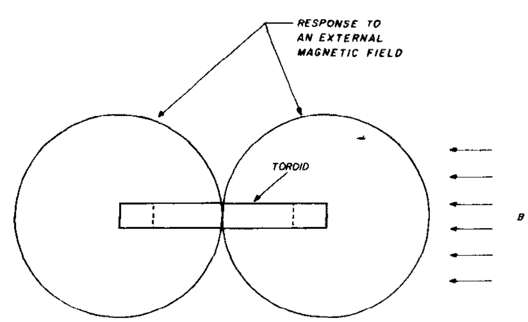

fig. 5. Response of the 1-turn effect of a conventionally wound toroid caused by the circumferential components to a magnetic field, B, parallel to the plane of the toroid.

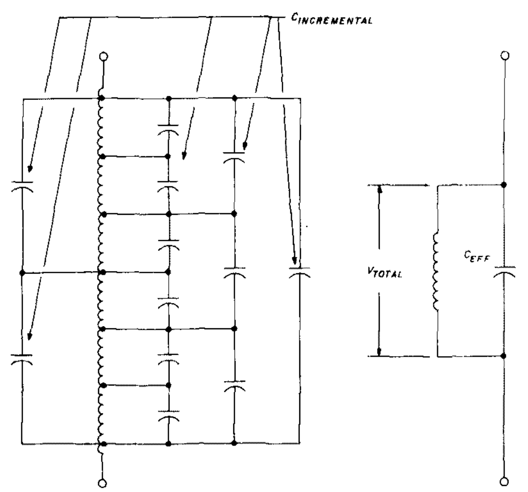

fig. 6. The effective capacitance, C_{EFF}, of an inductor stores the same amount of energy as the sum of the incremental capacitors, C_{INC}.

By adding one or two turns to the winding I could dramatically improve the performance at the low end of 80 meters, but the series resonant frequency occurred in the middle of the 10-meter band. Similarly, by removing a turn or two I could improve the isolation at 10 meters, but only at the expense of the 80-meter band.

In my opinion, the bandwidth performance of presently available baluns is limited by the following factors:

- State-of-the-art in presently available core materials — for example, higher permeabilities at the higher frequencies.
- The limitations imposed on the design by having to wind the core with a transmission line of a specified characteristic impedance rather than a single conductor as with the usual inductor.
- The need to handle an appreciable amount of power.

The reasons commercial baluns for Amateur applications are all wideband devices may include:

- Competition: no manufacturer can afford to offer a balun with less bandwidth than any other manufacturer.
- Lack of published specifications concerning isolation impedance, balance, and isolation make it impossible to tell how effective a given balun may be at the band edges.
- Economy: it is much less expensive for manufacturers and distributors alike to stock a single balun than several different "sizes."

For these reasons, I believe that if Amateurs want high-performance baluns optimized for particular bands, they will have to wind the balun themselves.

final comments

Physically, transmission-line baluns are very simple devices. This simplicity often causes some of the more subtle characteristics to be overlooked during design or use. Recognizing these subtleties may be the biggest problem in designing and using baluns.

reference

1. Walter Maxwell, W2DU, "Some Aspects of the Balun Problem," *QST*, March, 1983, page 38.
2. John J. Nagle, K4KJ, "Testing Baluns," *ham radio*, August, 1983, page 30.
3. Joe Reisert, "Simple and Efficient Broadband Balun," *ham radio*, September, 1978, page 12.

ham radio

short circuits

May 1985

digital HF radio

In "Digital HF Radio: A Sampling of Techniques" (April, 1985) the author's name was misspelled. "Dr. Ulrich L. Rhode," as it appears on page 18, should be corrected to read, "Dr. Ulrich L. Rohde." *ham radio* regrets the error.

May 1985

harmonic mixer

In **fig. 2** of K1ZJH's "Harmonic Mixer for VHF Signal Generation" (March, 1985, page 40), T1 should be identified as a T37-6 toroid available from Amidon Associates, Inc., 12033 Otsego Street, North Hollywood, California 91607.

May 1985

low-voltage power supplies

In fig. 9 of the March article, "Designing Low-Voltage Power Supplies," Q1, Q2, and Q3 are shown incorrectly wired. The base leads of Q1 and Q2 should be connected to the *collector* of Q3, not its base.

tilt-over conversion
of a
fixed antenna tower

Don't like climbing?
Try this.

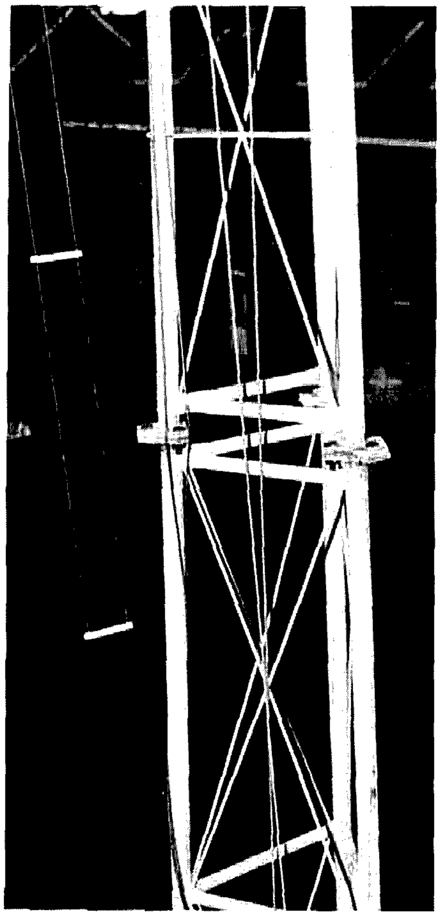

fig. 1. The tower before conversion. Hinge plates will be attached between the flanges connecting the upper and lower part of the tower.

By Dick Rollema, PAØSE, v.d. Marckstraat 5, 2352 RA Leiderdorp, Netherlands

Since 1973 I have been the proud owner of a self-supporting triangular steel lattice antenna tower. But because I'm afraid of heights, climbing this tower is out of the question. Changing antennas means engaging helpers, and even then it's not easy because the mast has an additional 10-foot (3 meter) long tube on top, making the rotator and antennas almost inaccessible.

It took a long time before I realized how I could convert my mast to a tilt-over. The 40-foot (12 meter) tower consists of two sections, each 20 feet (6 meters) long. (With the 10-foot, 3 meter tube on top, the total height comes to 50 feet, or 15 meters.) As shown in **fig. 1** the two sections are connected by three flanges with three bolts each; this division appeared to be the logical place for a hinge that would permit the upper section to pivot and be lowered to ground level.

The next problem to be solved was how to effect the tilt-over action by means of a winch and hoist line.

Sketches of the parts constructed by PAØDON are available from the author. Send a large self-addressed envelope; enclose three IRCs.

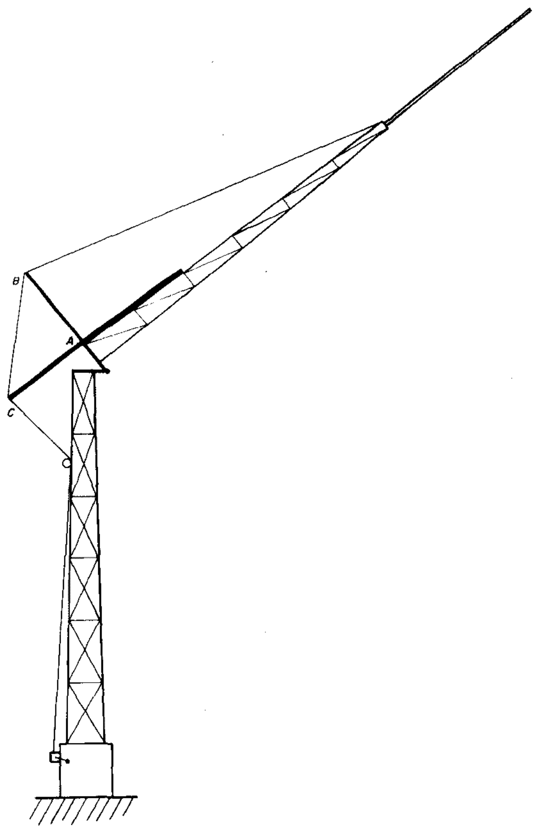

fig. 2. The initial plan for the tilt-over conversion was dropped because forces in outriggers *AB* and *AC* and the hoist line would be excessive.

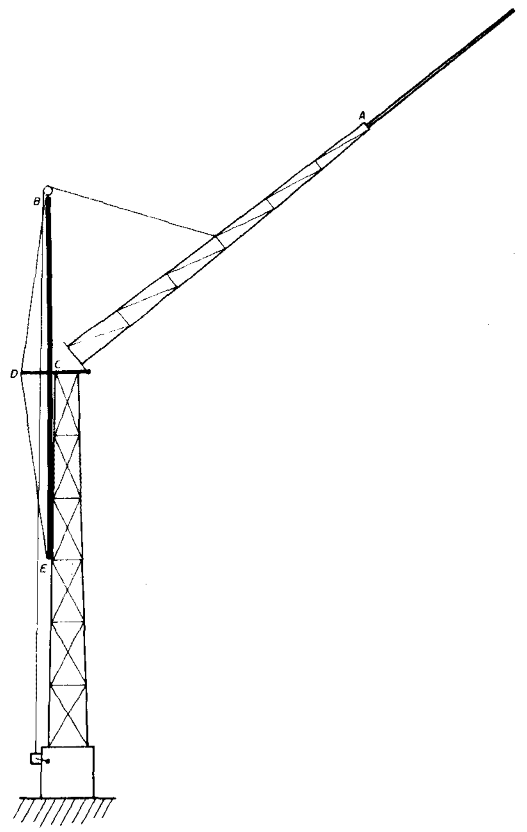

fig. 3. The final design: gin pole *BE* is 20 feet (6 meters) long. At *C* the gin pole is fastened to the lower hinge plate. The highest bending moment occurs at *A*.

I first thought of using a method such as the one shown in **fig. 2**, similar to the cranes used by building contractors. But with the tower positioned immediately behind our house, outriggers *AB* and *AC* would have to be short enough to clear the structure. This would lead to high tension in the hoist line and cause too much stress on the overall system. So the idea was dropped.

The breakthrough occurred when I visited PA0WCW, who has a similar tower that had been converted into a tilt-over. PA0WCW used two adjacent gin poles permanently fixed to the tower. A pulley is mounted on top of each pole. Two hoist lines in parallel run from the winch over the pulleys to a point halfway up the upper section of the mast.

My final design is shown in **fig. 3**. Here a single gin pole (*BE*) is fixed to the tower at the lower end *E* and in the middle at *C*. A few calculations showed that bending moment at *C* is so large that no pipe of acceptable size could endure this stress, so two stays designed to bear most of the load were introduced. These are fastened to the gin pole at both ends and pushed away from the middle of the pole by a triangular outrigger. **Figure 4** shows the finished system.

calculation is a must

Before working out a project such as this, it is absolutely essential to determine the forces exerted on the different parts. Calculating stresses in mechanical construction projects is definitely not part of my daily routine, but after consulting some textbooks — and with some professional advice from PA0TO — I managed to do the job. It was not very difficult after all. I quickly discovered that the most critical point is at *A* in **fig. 3** where the pipe top mast leaves the tower. The bending moment here should not exceed a certain maximum value set by the resistance moment and the tensile strength of the material.

My starting point is a pipe mast of 2-inch (50.8 mm) inner diameter with a resistance moment of 0.63 inches^3 (10.25 cm^3) that protrudes 10 feet (3 meters) out from the lattice mast. From a maximum admissible

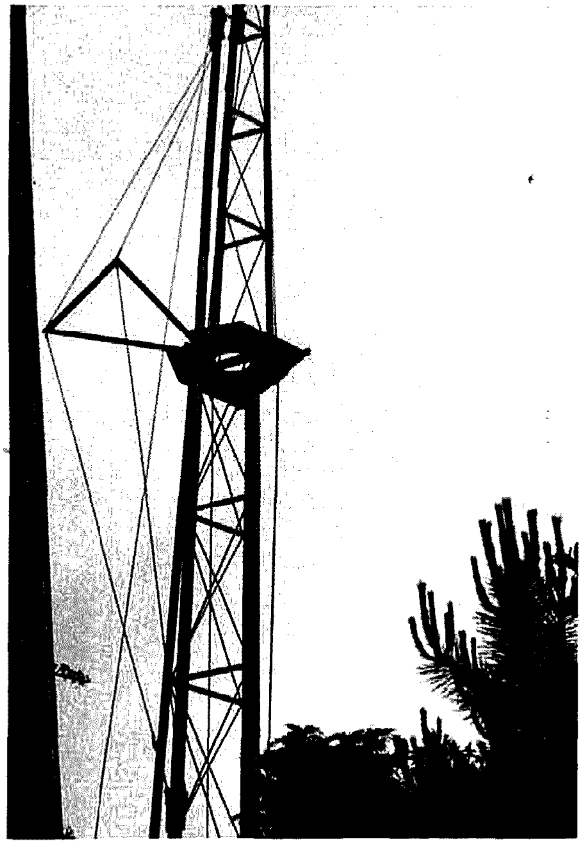

fig. 4. The completed structure.

stress of 23,205 pounds/inch^2 (160 N/mm^2) it follows that the maximum admissible lateral force at the top of the pipe is 123 pounds (547 N). This is the maximum wind load on antenna plus rotator, if we neglect the wind load on the pipe itself for just a moment.

A second limitation occurs when the mast has been tilted to a horizontal position. Again at *A*, a bending moment occurs as the result of the gravitational force on the top mast, rotator, and antenna. Taking into account the mass of the top mast itself, we find a maximum admissible mass of about 99 pounds (45 kg) for the rotator plus antenna. With these numbers, the forces in the different parts of the system can be computed. Rather than bore you with the results, only the maximum force in the hoist line will be mentioned: almost 1,034 pounds (4,600 N), which represents the static load. During tilting of the tower, shocks can easily occur that considerably increase the force in the line.

construction

As already mentioned, a hinge has been installed at the 20-foot (6 meter) level where the lower and upper parts of the tower are joined by flanges. The flanges are bolted to steel plates that carry the hinge at one edge.

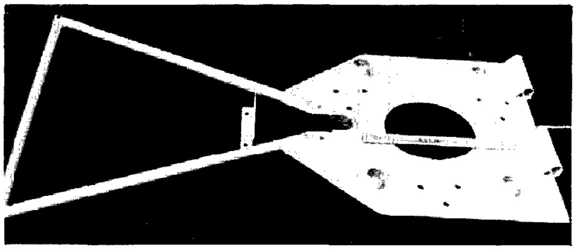

fig. 5. Lower hinge plate, made of 23/64-inch (9-mm) thick steel. The triangular outrigger is fabricated of 1-inch (25 mm) diameter steel tubing. The angle iron bracket secures the gin pole that is held in place in the half circular cutout. The central hole helps reduce the weight (64 pounds/29 kg).

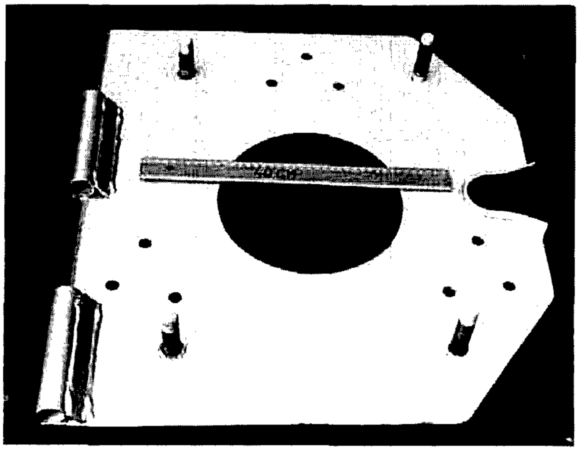

fig. 6. Upper hinge plate, made of 23/64-inch (9-mm) steel. The four bolts, welded on, keep the hinge plates together when the tower is upright. (Weight: 50 pounds/22.5 kg).

Figure 5 shows the lower plate and **fig. 6** the upper. The hinge is fabricated from a piece of 2-inch (50.8 mm) steel tube sawed into four parts. These were welded, two by two, to both plates. Extra strength was provided by welding strips of steel to both pipe and plates, as can be seen in **fig. 6**. The hinge pivots around a 1-inch (25-mm) diameter shaft with bushings at both ends. One is permanently fixed to the shaft; the other is fitted after mounting and secured by a bolt and nut that traverse the shaft and bushing.

In the resting position, the two hinge plates are held together by four 25/32-inch (20-mm) bolts that have been welded to the upper plate and pass through oblong holes in the lower plate. Pieces of tubing are welded around these holes to keep the plates at the proper distance.

Before tilting the tower, the four nuts must be removed from the bolts on the bottom side of the lower plate.

In my original design one plate lay directly on top of the other without any distance in between. To pre-

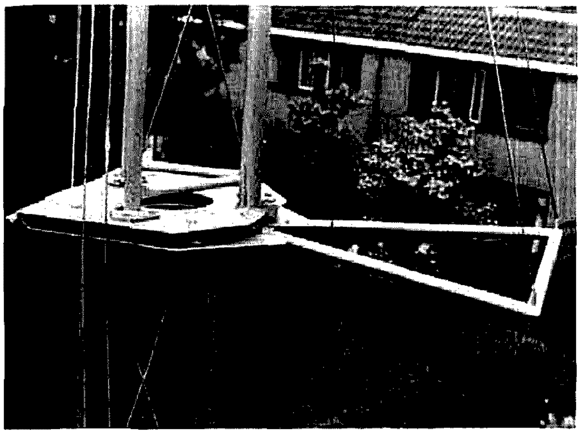

fig. 7. Tower in vertical position. The stainless-steel tubes that guide the stays through the outrigger can be seen at right.

vent the bolt heads in the upper and lower plates from interfering with each other, the upper part of the tower was rotated 60 degrees with respect to the lower part. This allowed the bolt heads of one plate to fall directly into holes drilled in the opposite plate. That 60-degree rotation has been maintained, though it is not necessary in the present construction. But while it presents neither advantage nor disadvantage, it does lead to an optical illusion: from whatever side the tower is viewed, it always seems that the center lines of the upper and lower part do not coincide.

The tower is hinged so that the top mast can be reached from the roof of the barn in the back of our garden, so it is easy to work on rotator or antenna. **Figure 7** shows the tower and hinge in its normal position. **Figure 8** shows it tilted over about 90 degrees.

The gin pole is a seamless steel tube 20 feet (6 meters) long and 2.5 inches (60.3 mm) outer diameter; its inner diameter is 2 inches (53 mm). The lower end is supported by the parts shown in the lower section of **fig. 9**. **Figure 10** shows the lower end in position on the tower.

The legs of the tower have a wall thickness of only 1/16 inch (1.5 mm). To spread the forces exerted by the U-bolts, my friend PAØDON fitted half-cylinder-shaped shells between the U-bolts and the tower legs. On the angle-iron bracket, a piece of tubing was welded over which the gin pole was placed. After mounting, a hole was drilled for a long bolt that pierces the gin pole and prevents turning.

On top of the gin pole a pulley was mounted by means of the method shown in **figs. 11** and **12**. As shown, precautions have been taken above and behind the pulley to prevent the hoist line from running off. The pulley, made of Novotex,™ has an outer diameter of 4.7 inches (120 mm). It rotates around a piece of steel tubing of 1.3 inches (32 mm) diameter. The steel strips that hold the pulley on top of the gin

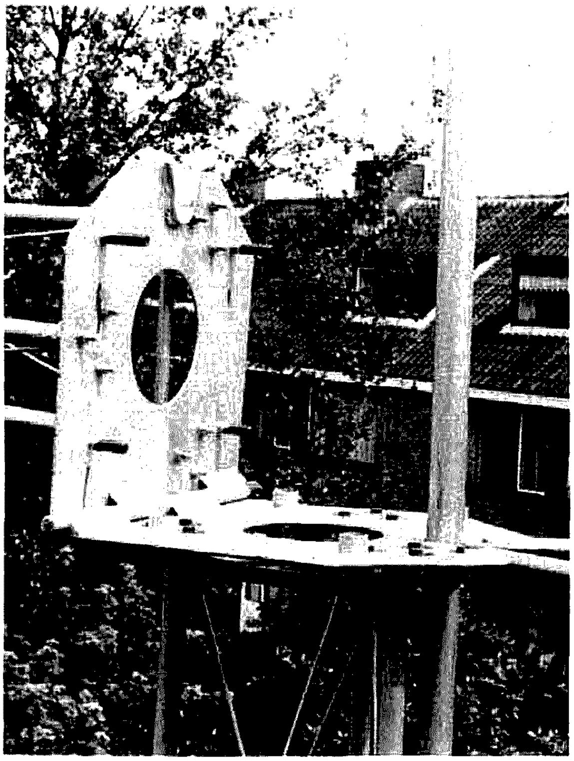

fig. 8. Tower tilted 90 degrees.

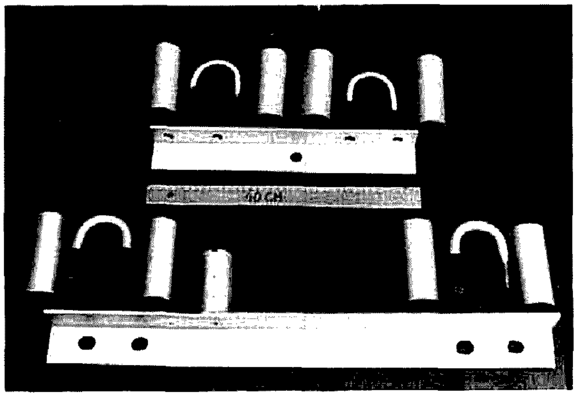

fig. 9. The parts that connect the hoist line to the upper tower section are shown above; the support for the gin pole is shown below.

pole also carry bolts for the stays. At the lower end the stays are connected with stainless steel turnbuckles with a breaking strength of 4,271 pounds (19,000 N). The lower hinge plate carries a triangular outrigger that keeps the stays at the proper distance from the middle of the gin pole (**fig. 7**). To prevent chafing, stainless steel tubes have been inserted in the outrigger that guides the stays. A nylon bushing inside the tubes provides extra protection.

Stays and hoist line consist of stranded stainless

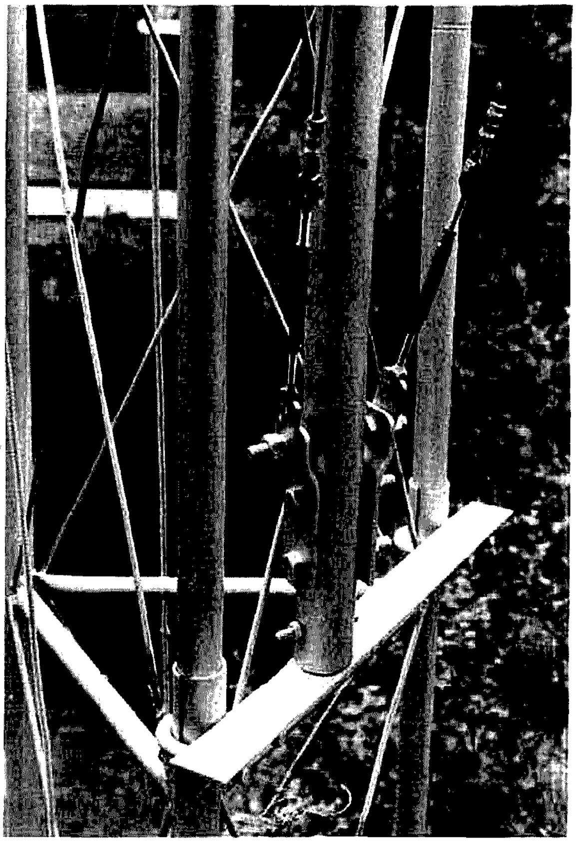

fig. 10. The lower end of gin pole, with stays.

steel cable measuring 0.2 inch (5 mm) in diameter and having a breaking strength of 3,147 pounds (14,000 N). The hoist line ends in a winch with a capacity of 1,200 pounds (5,338 N) and a reduction of 4.1:1. (If I were to build this system again I would select a larger reduction). The tower rests on a concrete block, and the winch is fitted to this block by three "chemical" anchors measuring 0.31 inch (8 mm) (**fig. 13**). The upper end of the hoist line ends halfway up the top section of the tower. A bracket of angle iron is shown in the upper half of **fig. 9**. This bracket is mounted to the tower legs with U-bolts, and half-cylinder shells are installed to distribute the pressure more evenly. The hole in the bracket is for a 0.625-inch (16 mm) bolt. (The hoist line is fitted to this bolt with parts purchased at a marine shop; I am, unfortunately, unable to translate their names into English.)

The hinge plates and all parts visible in **figs. 9** and **11** have been treated with zinc epoxy followed by two coats of aluminum paint. The gin pole, galvanized by the manufacturer, was given a coat of rust-preventive primer and two coats of aluminum paint. Although the tower was also galvanized, some traces of rust were visible at the upper end, so the whole thing was also treated with rust-preventer and aluminum paint.

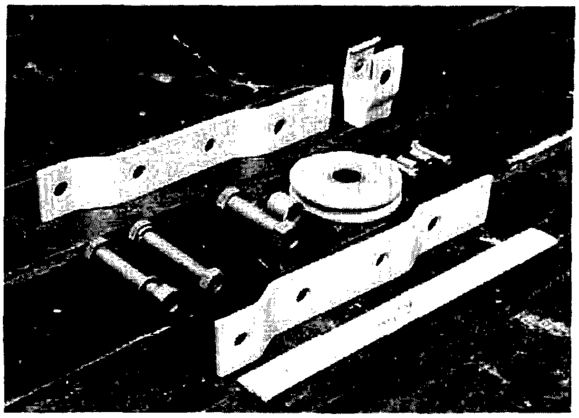

fig. 11. These parts support the pulley on top of the gin pole.

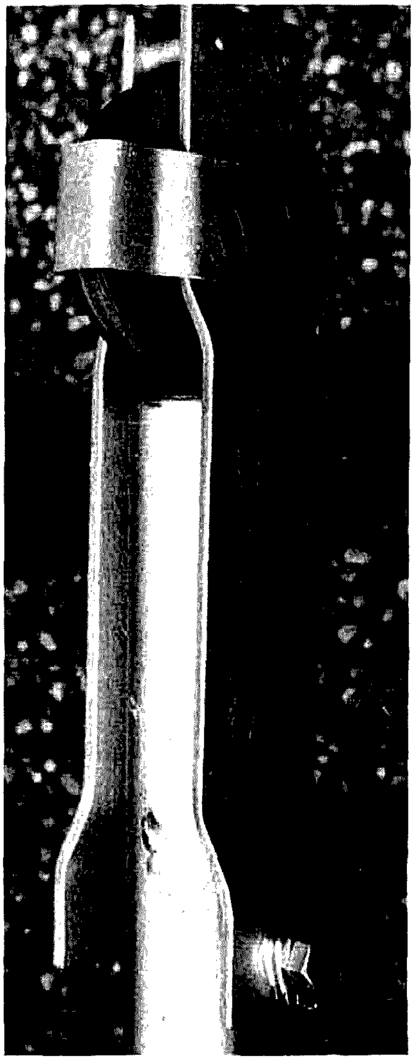

fig. 12. Pulley in position.

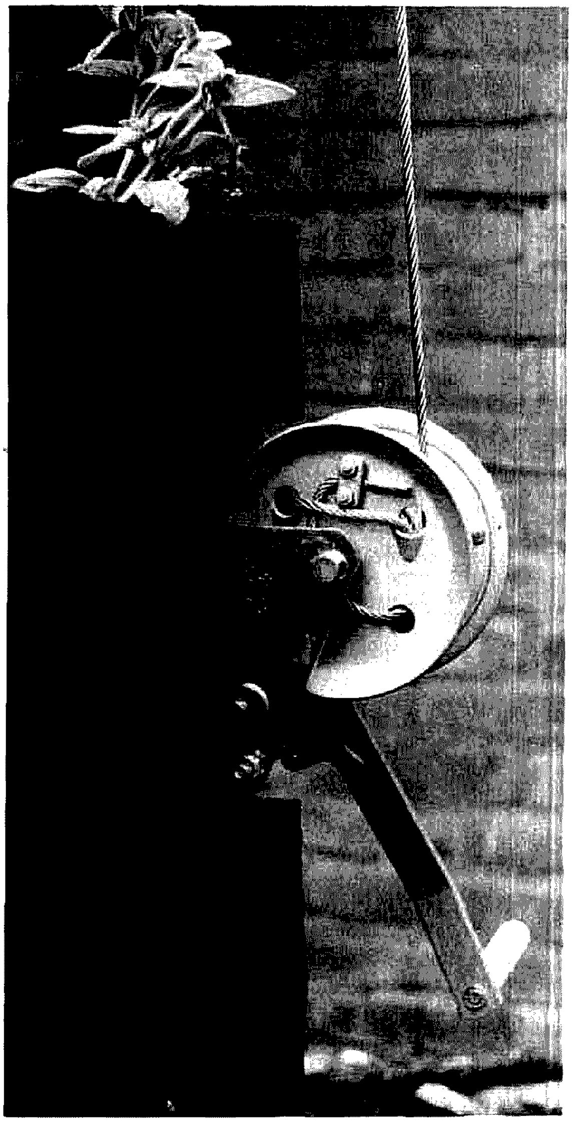

fig. 13. The winch is fastened to the concrete block that serves as base for the tower. Concrete block is concealed under a wooden cover for aesthetic purposes.

assembly

The first phase of the operation took place in early December, 1983, when the temperature was just below freezing. Our object was to take down the upper part of the tower, the top tube mast, the rotator, and the 2-meter beam. The gin pole was fitted temporarily to the tower. Because the hinge plates were not yet in place, the gin pole was tied to the upper end of the lower section with rope. The hoist line was also fitted. Using the winch, we gently lowered the upper part of the tower to ground. Now the holes in the flanges could be transferred to a template to be used later in drilling the holes in the hinge plates.

fig. 14. High level work: Jos (PA3ACJ) and SWLs Bert and Ed.

Final assembly and rigging of the tilt-over conversion occurred in mid-May, 1984, under more favorable weather conditions. Everything fit exactly, and it was a real pleasure to see the converted tower come together step by step (**fig. 14**).

acknowledgement

My sincere thanks to Ber van Dongeren, PA0DON, for his superb realization of my design. No operation like this can be performed without helpers: Piet de Bondt, PA3BGP; Jos Disselhorst, PA3ACJ; Ton Verberne, PA2ABV; Ben van Capel, PE1KCG; Gerrit van Zwam, PE1KAX, and SWLs Jo Chin-Chan-Sen, Bert Kraan and Ed Wassenburg helped disassemble the original tower and erect the new one.

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AWARDS DIRECTORY $1.00 brings information. Refundable. Updated monthly. Computerized. Worldwide. List your awards free. Over 1000 now in files. KB0ZP. RR#1B, Cumming, IA 50061.

EQUIPMENT SALE. HAL DS2000 KSR RTTY/CW keyboard $100. Panasonic 9" monitor $80. Collins SM280 microphone $60. Six good used 4-400A's $20. each. TR3/AC3/MS4 $120. 1.6 kHz S-Line filter $40, 4-1000A and parts for amp. Mike Palmer, K5FZ, 16707 Creeksouth, Houston, Texas 77068. (713) 444-7737

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QST Magazines. 271 of them about 200 lbs. 1930, 1933 through 1940 solid, 1952 through 1965 solid. Best offer for the whole works only. W2LTJ.

FREE CATALOG of books for electronics professionals and hobbyists. Send SASE to Technical Advice, P.O. Box 4757, Englewood, Colorado 80155.

NEW low frequency products by K1RGO. L-101 VLF converter (2 kHz to 500 kHz), AM broadcast rejection 100 dB, i-I rejection 130 dB. L-201 VLF fixed tuned preamplifier (2 kHz to 500 kHz), 25 dB gain (adjustable). Free brochure. LF Engineering Co., 17 Jaffry Rd., East Haven, CT 06512.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

CHASSIS and cabinet kits. SASE K3IWK.

IMRA, International Mission Radio Association, helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, NY 10538.

WANTED: Old microphones, remote mixers other misc related items. All pre-1935. Box Paquette, 107 E. National Avenue, Milwaukee, WI 53204.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 09224 (213) 774-1255.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details. W8UUR, 1025 Wildwood Road, Quincy, MI 49062.

Coming Events

ACTIVITIES

"Places to go..."

INDIANA: Hamfest and Computer Show, Sunday, June 9. Sponsored by the Muncie Area Amateur Radio Club. 8 AM to 3 PM. Rain or Shine. Delaware County Fairgrounds. Tickets $2.00 advance, $3.00 at door. Large Flea Market. Overnight camping on grounds. Hookups $5.00. Talk in on 146.13/.73. For tickets, table reservations or information: SASE to MAARC, PO Box 2102, Muncie, IN 47302.

OHIO: The Athens County Amateur Radio Association's sixth annual Hamfest and Computerfest. Sunday, May 19, City Recreation Center, East State Street, Athens. 8 AM to 3 PM. Admission $3.00. Free paved outdoor tailgating or bring your own tables. Indoor space by advanced registration only. Contact Joe Follrod, NE8R, 15 Roy Avenue, The Plains, OH 45780. (614) 797-4874. This year there will be all level license exams. Mail completed FCC Form 610 and $4.00 check payable to ARRL/VEC to John Cornwell, NC8V, exam coordinator, 101 Coventry Lane, Athens, OH 45701. Forms must be received by April 19. Talk in on club repeater 146.34/.94. For information: Carl J. Denbow, KA8JXG, Chairman, ACARA Hamfest Committee, 63 Morris Ave., Athens, Ohio 45701.

LOUISIANA: The Central Louisiana ARC (CLARC) will sponsor a Hamfest, Saturday and Sunday, June 8 and 9, Bolton Avenue Community Center, 315 Bolton Avenue, Alexandria. Free admission. Swap tables available. VEC exams. Talk in on 147.93/33 or 146.04/64. For more information: CLARC, PO Box 7772, Alexandria, LA 71306.

GEORGIA: The John Ross Amateur Radio Club's annual Hamfest, June 8 and 9. New location — Lakeview — Ft. Oglethorpe High School, Ft. Oglethorpe on Hwy. 2A. Air conditioned dealer spaces. Indoor/outdoor flea market. Tables available. FCC exams both days. Refreshments and plenty of free parking. Talk in on 145.35/down. For reservations and information: JRARC. PO Box 853, Rossville, GA 30741. (404) 861-5610.

MICHIGAN: The Chelsea Swap and Shop, Sunday, June 2, Chelsea Fairgrounds. Gates open for sellers at 5 AM. General public 8 AM to 2 PM. Donation $2.50 advance or $3.00 at gate. Children under 12 and non ham spouses admitted free. Talk in on 146.520 simplex and 147.255 Chelsea repeater. For information: William Altenberndt, 3132 Timberline, Jackson, MI 49201.

MINNESOTA: The North Area Repeater Association will sponsor the state's largest swapfest and exposition, Saturday, June 1, Minnesota State Fairgrounds in St. Paul. Free overnight parking for self-contained campers May 31. Exhibits, dealers, giant flea market. Amateur license exams given. Admission $4 in advance, $5 at Fair. Talk in on 25/85 or 16/76 repeaters. For tickets, information: Amateur Fair, PO Box 857, Hopkins, MN 55343. (612) 566-4000.

MARYLAND: The Maryland FM Association's annual Hamfest, Sunday, May 26, Howard County Fairgrounds, West Friendship, I-70, 30 miles west of Baltimore. Gate opens 8 AM to 4 PM. Inside tables by advance registration $6.00. At the door $10.00 if available. Donation $3.00. Talk in on 146.16/76, 222.16/223.76, 449.1/444.1. For tables or information: Craig Rockenbauch, WA3TID, 429 Severnside Drive, Severna Park, MD 21146. (301) 987-6042 (6-10 PM).

GEORGIA: The Anderson, Hartwell and Toccoa Amateur Radio Clubs will hold the 6th annual Lake Hartwell Hamfest, May 18 and 19, Lake Hartwell Group Camp, Highway 29, Hartwell. Free admission, free camping, and free flea market space. Left footed CW contest, horseshoes, fishing, swimming and camping available on site. Campgrounds open 6 PM Friday. Talk in on 146.19/.79, 147.93/.33 and 146.895/.295. For further information: Ray Pettit, WB4ZLG, Rt. #1, Dooley Drive, Toccoa, GA 30577.

MARYLAND: The Frederick Amateur Radio Club's 8th annual Hamfest, June 16, Frederick Fairgrounds. Admission $3.00. YL's and children free. 8 AM to 4 PM. Tailgaters $2.00 extra. Gates open for exhibitors 8 PM June 15. Overnight parking welcomed. Tables $10.00 first, $5.00 each extra. For information: Jim Kasunic, KA3LPC, 9419 Highlander Ct., Walkersville, MD 21793.

ILLINOIS: The Chicago Amateur Radio Club (CARC) is sponsoring a Mini Hamfest, Wednesday, May 15, 6 to 10 PM, Edgebrook Golf Course Fieldhouse, 5900 N. Central Avenue, Chicago. (North of Elston Avenue, south of Devon Avenue). For more information: (312) 545-3622.

NEW HAMPSHIRE: The 11th annual Eastern VHF/UHF Conference, May 17-19, Rivier College, Nashua. Friday night hospitality room. Saturday tech talks by well-known VHFers. Pre-registration $13.50 to David Knight, KA1DT, 15 Oakdale Avenue, Nashua, NH 03062 before May 5. Registration at the door $20.00. Saturday night banquet $14.00 payable before May 5. Dormatory housing available, $16.50 per night single, $28.00 per night double, includes buffet breakfast. Include payment at time of pre-registration. Make checks payable to Eastern VHF/UHF Conference. There are also numerous hotels and motels nearby. For information: Lewis D. Collins, W1GXT, 10 Marshall Terrace, Wayland, MA 01778. (617) 358-2854 (6-10 PM).

OHIO: The 5th annual Columbus Hamfest sponsored by the Battelle Amateur Radio Club, Sunday, June 2, 8 AM to 3 PM. Ganyard Building on Franklin County Fairgrounds. Admission $2 advance and $3 at the door. Advance tables $3; at the door $4. Talk in on 146.37/97. For information Bill, W8LLU (614) 261-7053 or Kevin, WA8OHI (614) 766-5313. Advance sales SASE to Bill Welch, W8LLU, 396 Brevoort Rd., Columbus, Ohio 43214.

OHIO: The Sandusky-Ottaway Counties annual Hamfest, May 19, 8 AM to 3 PM. Ottawa County Fairgrounds, St. Rt. 163 east of Oak Harbor. Donation $3.00 at gate; $2.50 advance. Tables $3.00. Gate opens 6 AM for dealers and 8 AM for

general public. Trunk sale (donation 50¢), plenty of parking, food and refreshments. To reserve tables: Raymond Kruse, K8IDA, 18980 W. S.R. 51, Elmore, Ohio 43416. (419) 862-2619 (8 AM - 5 PM).

PENNSYLVANIA: The Warmister ARC Hamfest, May 19, Middletown Grange Fairgrounds, Penns Park Rd., Wrightstown (about 15 miles north of Phila.) Admission $3.00. XYLs and children free. Indoor space with table and elec $5.00. Tailgating space $5.00. Gates open 7 AM. Talk in on 147.69/09 and 147.52. Computer hardware/software vendors welcome. For information: Bill Cusick, W3GJC, Apt. 706, Garner House, Hatboro, PA 19040. (215) 441-8048.

NEW YORK — Long Island: The Suffolk County Radio Club's Indoor/Outdoor Electronic Flea Market, Sunday, May 5, 8 AM to 3 PM, Republic Lodge No. 1987, 585 Broadhollow Road (Rt. 110) Melville. Refreshments and plenty of free parking. General admission $2.00. Wives and children under 12 free. Indoor seller's tables $7.00. Outdoor space $5.00 includes one free admission. Talk in on 144.61/145.21 and 146.52. For information: Richard Tygar, AC2P. (516) 643-5956 evenings.

NEVADA: The 10th International Convention of the YLRL will be held June 20, 21, 22 and 23 at the Sahara Hotel, Las Vegas. Deluxe accommodations and RV parking available at reasonable rates. Activities include a Hoover Dam tour, Lake Mead cruise, desert tour, gala stage show, cocktail party, luncheon buffet and awards banquet plus a DX YL show, slide shows and business meetings. A convention station will operate on 14,288 kHz and other frequencies. Registration forms are to be found in recent issues of YLRL's "Harmonics". For information packet send business SASE with 2 oz. of postage to Jan Weaver, N7YL, 2195 East Camero Avenue, Las Vegas, NV 89123.

NEW JERSEY: The Tri-County Radio Association's Hamfest, Sunday, May 19, rain or shine, Passaic Valley Community Center off Valley Road, Stirling. 9 AM to 4 PM. Indoors, free parking, refreshments. Tables $10. Registration $2.50. Reserved tailgating only. All reservations to: Dick Franklin, W2EUF, PO Box 182, Westfield, NJ 07090. (210) 232-5955 or 270-3193.

SOUTH DAKOTA: The Black Hills Amateur Club is celebrating its 50th anniversary by sponsoring the 1985 ARRL Dakota Division Convention, July 5, 6, 7, Howard Johnson's (I-90, exit 59) Rapid City. Registration Friday from 4 to 7:30 PM followed by a social hour. Saturday exhibits, registration, indoor flea market (free tables while they last) ladies' activities and fun for the whole family. VE exams (limited walk-ins) 1 PM. Banquet at 7 PM. Sunday brunch at 8:30 followed by the ARRL forum. Exhibits will be open. Pre-registration & banquet $18.00. Pre-registration only $6.50. Must be made prior to 6/10/85. Additional banquet tickets $12.50. Sunday brunch $6.75, children 12 and under $3.75. Registration after 6/10/85, $7.50. Make check payable to Black Hills ARC, c/o Gene F. Bauer, KXØU, 713 Blaine Avenue, Rapid City, SD 57701. Confirmation by return mail.

WISCONSIN: The Central Wisconsin Radio Amateur's annual Swapfest/Family Picnic, June 16, Bukolt Park, Stevens Point. Swapfest will feature volunteer exams (by advanced registration only), dealers, food, refreshments and fellowship. Tables and tailgating $2.00. Talk in on 146.07/67 and .385/985. For further information: Jim Benak, KA9ACE, 1775 Strongs Avenue, Stevens Point, WI 54481. To register for exams (by May 16): Gene Santoski, K9UTQ, 1220 — 18th Street South, Wisconsin Rapids, WI 54494.

KENTUCKY: The Armored Force Amateur Radio Nationwide Emergency Team (A FAR NET) will hold an Eyeball Bivouac, June 7 - 9, at the Best Western Gold Vault Inn in Radcliff, just outside Fort Knox. Campers will use the Fort Knox campground at Camp Carlson. On Friday, June 7, there will be an informal picnic. Saturday, the group will take a guided bus tour of Fort Knox ending at the Patton Museum. Saturday evening's cocktail party is followed by dinner at 8 PM. There will be a farewell breakfast Sunday morning. For additional information contact Carl Quickmire, WB4UBS, 6341 Cloverdale Drive, Columbia, SC 29209. Any Amateur Radio operator who has ever served with or been assigned to an armored unit of the U.S. armed forces or its allies is eligible for membership in A FAR NET. Family members holding an Amatuer license are eligible for associate memberships. Anyone interested in joining this unique organization can obtain information and an application from Harry B. Thomsen, W2PJH, 348 Jefferson Avenue, Apt. 15, Canandaigua, NY 11424. Please include business SASE.

PENNSYLVANIA: The 31st annual Breeze Shooters Hamfest, Sunday, June 2, 9 AM to 5 PM, White Swan Amusement Park, Rt. 60 near Greater Pittsburgh International Airport. Free admission and flea market. Family amusement park. Registration $2.00, 3 for $5.00 or 7 for $10.00. Vendors tables by advance registration. Talk in on 146.28/.88 or 29.000 MHz. For information: John Colbert, K3SDL, 1831 Highland Ave., Irwin, PA 15642. (412) 863-5167 evenings only.

COLORADO: The Longmont Amateur Radio Club's annual Boulder Spring Hamfest, Sunday, May 5, 9 AM to 2 PM rain or shine, Colorado National Guard Armory, 4750 North Broadway, Boulder. Donation $3 per family. No sellers charge but bring own tables . . . Hamswap, plus tech demonstrations and seminars. Food and drink available. Talk in on 146.16/.76 and 146.52. For information: William Currie, WDØEHJ, 1232 East Fourth Ave., Longmont, CO 80501. (303) 776-2829.

NEW YORK: Long Island Hamfair sponsored by LIMARC, Sunday, June 9, 9 AM to 4 PM, Electricians Hall, 41 Pinelawn Road, Melville, Long Island. General admission $3.00. $2.00 after 1 PM. Table space in advance from Hank Wener, WB2ALW, 53 Sherrard St., East Hills, NY 11577-1712. 4 x 6' table space $10.00 or your own for $6.00. Contact Hank at (516) 484-4322 evenings to 11:30 PM.

INDIANA: The 39th Annual Hamfest sponsored by the Wabash Valley ARA, Sunday, June 2, Vigo County Fairgrounds, Terre Haute. For information SASE to WVARA, PO Box 81, Terre Haute, In 47808.

CONNECTICUT: Flea Market, sponsored by the Newington Amateur Radio League, June 9, Newington High School, 9 AM to 2 PM. Admission $2.00. Tailgating $5.00. Talk in on 146.52 or W1AW/R 145.450, 224.840.

NEW YORK: The Rome Radio Club's 33rd annual Ham Family Day, Sunday, June 2, Beck's Grove in Rome. Games, contests and large Flea Market. Good food and beverages available throughout the day. Educational and scientific presentations. The day will end with dinner and the presentation of "Ham of the Year Award". For further information: Rome Radio Club, PO Box 721, Rome, NY 13440.

NEW YORK: The 26th annual Southern Tier Amateur Radio club's Hamfest, Saturday, May 4, Treadway Inn, Owego. Flea market opens 8 AM. Vendor displays and sales. Dinner at 6:30 PM by advance tickets only. Talk in on 22/82, 16/76 and 146.52. For further information SASE to KF2X, RD #1, Box 144, Vestal, NY 13850.

NEW YORK: The Antique Radio Club of America will hold an international convention at Niagara Falls, June 5-8. The club has 1000 members who collect and restore antique wireless and radio equipment and who study and record the history of early radio. For information on the convention or membership in ARCA please write NFWA, PO Box 68, Central Park Station, Buffalo, NY 14215.

WASHINGTON: The Yakima Amateur Radio Club will hold the Central Washington State Hamfest, May 18 and 19, Hobby Building at Central Washington State Fairgrounds, Yakima. Saturday 9 to 5 with lunch available. Sunday 7 to 2 with breakfast and lunch available. Registration $4.00 advance, $5.00 at the door. Free swap and shop with plenty of tables. Talk in on 146.01/61 and 146.34/94. For pre-registration contact Tom Plaisance, PO Box 9211, Yakima, WA 98909.

OPERATING EVENTS

"Things to do..."

MAY 5 AND 5: The Mason County ARC will operate commemorative stations KB7MJ and W7KTI on SSB, KN7D RTTY and K7UAR on Packet to celebrate the Shelton, Washington, Centennial. Certificates will be exchanged for a QSL card and 9 x 12 SASE. Send to Loren Mercer, KA7GSV, 2213 Olympic Hwy. North, Shelton, WA 98584.

JUNE 2: SRRC Hamfest, Princeton, Illinois. Plans include FCC/VEC exams. Registrations $2.50 before May 20. $3.00 June 2. For advance registrations and/or complete information, furnish a long SASE to Starved Rock Radio Club, W9MKS, RFD #1, Box 171, Oglesby, Illinois 61348. (815) 667-4614.

JUNE 1 AND 2: The Southside Amateur Radio Club will operate station NØEWP in honor of President Harry Truman's 101st birthday. The station will operate near the old Truman farm home in Grandview, MO during the annual "Harry's Heydays" celebration. For a commemorative certificate send 9 x 12 SASE with 33¢ postage to Southside ARC, PO Box 412, Grandview, MO 64030.

MAY 4: The Sand Hills Amateur Radio Club will operate KZØM during a DXpedition to Moscow, KS to commemorate May Day. For a special OSL SASE via Box 88, Moscow, KS 67952.

MAY 25: The Bay Area Amateur Radio Society, Pasadena, Maryland, will operate KM3I and KA3HKB to commemorate "Samuel F.B. Morse Day". For a special certificate send QSL and large SASE to B.A.A.R.S., PO Box 805, Pasadena, MD 21122-0805.

CQ CONTEST: VHF'ers please note! The first annual CQ World Wide VHF WPX Contest is July 20-22, 50 thru 1296 MHz. For details, logsheets, etc., write to SCORE, PO Box 1161, Denville, NJ 07834 or to CQ Magazine. We need your entry to make this a success.

MAY 18: ARRL/VEC License Exam at 9 AM, United Community Services Building (Red Cross), 39 North Park Street, Mansfield, Ohio. Send SASE, 610 and check for $4 payable to ARRL/VEC to Peggy Boyle, KC8NH, 1464 Marion Avenue Road, Mansfield, Ohio 44906.

MAY 18 AND 19: The Amateurs of Lima and Allen County, Ohio will commemorate the discovery of oil in Lima. Stations participating will sign /OIL and operate in the Novice, Technician or General portions of the bands. For a certificate SASE to Northwest Ohio ARC, PO Box 211, Lima, Ohio 45801.

JUNE 1: The Wireless Institute of Northern Ohio (W.I.N.O.) will be operating a special event station from an actual winery in Madison, Ohio to commemorate Ohio Wine Month. Listen for KO8O Saturday evening 7 and 11 PM EDST and Sunday 11 AM to 3 PM EDST. A special 8½ x 11 QSL certificate will be available from KO8O — WINO Weekend, 7126 Andover Drive, Mentor, Ohio 44060.

JUNE 8: Tri-City ARC will operate W7VPA from Ice Harbor Dam the highest lift navigable locks in the United States. Operation will be from 1700 to 2400Z. For a special QSL SASE to W7VPA, PO Box 73, Richland, WA 99352.

MAY 10: The Owensboro ARC will operate K4HY from 0000Z to 0530Z May 11 to celebrate their International BBQ Festival. Certificate for SASE via N4EKG, 1615 East 23rd Street, Owensboro, KY 42301.

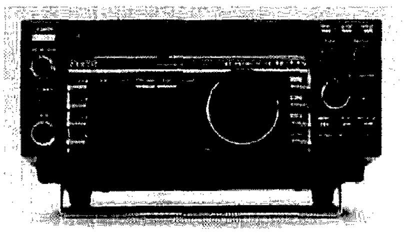
NEW products

new IC-735 HF transceiver

ICOM has announced the IC-735 ultra compact all-ham band HF transceiver and general coverage receiver, said to be the most compact and advanced HF SSB unit on the market. Measuring only 3.7 inches high by 9.5 inches wide by 9 inches deep, the IC-735 is suited for mobile, marine, or base station operation.

To enhance receiver performance, the IC-735 has built-in receiver attenuator and preamp. It also has a 105-dB dynamic range and a new low-noise phase locked loop for rock-solid reception.

The IC-735 features a large LCD readout and conveniently located controls that allow simple operation, even in the mobile environment. VOX controls, mic gain, and other infrequently changed controls are kept out of sight behind a hatch cover on the front panel of the radio, but are immediately accessible.



Other standard features included at no additional cost are: built-in FM, a 500-Hz CW filter (FL-32), an electronic CW keyer, an HM-12 scanning mic, FM/CW/LSB/USB/AM (TX and RX). 12 tunable memories with lithium memory backup, and both program and memory scan. Adjustable AGC, automatic SSB selection by band, an RF speech processor, and 12 volt operation are also standard. Output power is continuously adjustable up to 100 watts.

A new line of accessories, including the AT-120 automatic programmable antenna tuner and the PS-55 power supply, will also be available. The IC-735 is also compatible with most of ICOM's existing line of HF accessories.

The IC-735 will be available in May, 1985, and pricing will be announced the end of April. The IC-735 will be displayed at the Dayton Hamfest.

For additional information, contact ICOM America, Inc., 2380-116th Avenue, N. E., Bellevue, Washington 98004.

Circle #301 on Reader Service Card.

PV voltage regulators

Photocomm, Inc., of Scottsdale, Arizona, has announced new design features on its Power Control Series voltage regulators for photovoltaic applications.

A removable printed circuit board has been incorporated to simplify servicing and repairs; spare boards can also be purchased separately. These regulators now have a 30-amp load handling capability that allows for up to 12 40-watt solar modules in parallel to be controlled simultaneously. This benefit helps to lower overall system cost by using fewer regulators.

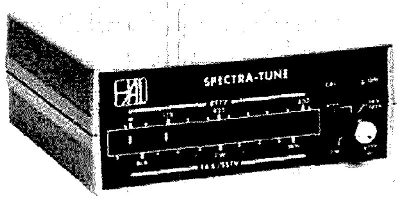
Standard features include temperature compensation, low voltage disconnect, analog meter package, transient protection and adjustable voltage setpoints. The controller is housed in a NEMA 3R weather-proof enclosure. 12 and 24 volt units are available off the shelf; other voltage units (36, 48, and 120 volts) can be provided upon request. Special options include Array Power Diversion, Manual Array Switching and a Severe Environment Package.

For further information, contact Photocomm, Inc., 7735 E. Redfield Road, Scottsdale, Arizona 85260.

Circle #302 on Reader Service Card.

multi-mode tuning indicator

Get on frequency quickly and accurately with the new HAL SPT-1 "Spectra-Tune" multi-mode tuning indicator. The SPT-1 may be used in the following modes of operation: RTTY (high tones) and RTTY (low tones), Fascimile (FAX), Slow-Scan Television (SSTV) and Morse Code (CW).



The frequency spectra of the received signal is displayed on a calibrated linear 40-segment red LED bar graph. The front panel calibration control allows the selection of which 1000 Hz range

NEW products

from 300 Hz to 3000 Hz is to be displayed. Best of all, hookup is very simple — and audio input line from the receiver/transceiver and a +12 VDC line are all that's required for operation. The SPT-1 comes with a one-year limited warranty.

The SPT-1 Spectra Tune is priced at $169.00.

For more information, contact HAL Communications Corporation, P.O. Box 365, Urbana, Illinois 61801.

Circle #303 on Reader Service Card.

KENPRO rotator products

The KENPRO brand line of rotators and accessories — including the KR-400 and KR-500 pair of rotators for satellite and space communications — are now available from Encomm, Inc. Also available are the KR-600RC medium-duty and the KR-2000RC heavy-duty units for use on the U.S.A. voltage system.

For further details, contact Encomm, Inc., Suite 800, 2000 Avenue 'G', Plano, Texas 75074.

Circle #304 on Reader Service Card.

VHF and UHF transmitter and receiver modules

Hamtronics, Inc. recently announced commercial versions of its popular high-band and UHF FM transmitter and receiver modules, FCC Type Accepted under parts 15, 22, 74, and 90. The TA-51 VHF transmitter and the TA-451 UHF transmitter are each capable of operating at 2 watts continuous duty and up to 3 watts intermittent service. The companion R144 and R451 receivers feature high sensitivity and superior front-end and IF selectivity, automatic frequency control, and hysteresis squelch to lock onto drifting or fading signals.

The Hamtronics® transmitter and receiver modules are supplied as printed circuit board assemblies that can be housed in a variety of ways and adapted to many applications. The low-cost units are suitable for use in voice and data links to replace phone lines, for remote control, for wireless alarm systems, for telemetry, for monitor receivers, and for many other applications.

Other Hamtronics® products of interest include "202" type FSK modulators and demodulators to provide digital data operation with FM transmitters and receivers, low-noise receiver preamps, 806-MHz band scanner converters, VHF AM receivers, repeater CORs, CWIDs, simplex and duplex autopatches, DTMF control and many other products related to VHF and UHF transmission and reception.

For a copy of Hamtronics' new 40-page catalog, contact Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535. (For overseas mailing, send $2 or 4 IRCs.)

Circle #118 on Reader Service Card.

THE GUERRI REPORT

Ernie Guerri W6MGI

a busy signal from space

For over twenty-five years various nations have been launching satellites of one type or another. Many of the early satellites were experimental in nature, primarily intended to develop and test space hardware. However, the past fifteen years have been devoted to the more utilitarian aspects of space. Because our only link to satellites has been through RF paths, the amount of the spectrum devoted to this purpose has grown dramatically.

The geosynchronous equatorial orbits are particularly popular. This is because satellites placed in this position appear to remain in one place all the time, and can illuminate a large portion of the earth's surface. Because of the desirability of this location, it has become imperative that there be global cooperation with respect to the placement of satellites in the general equatorial area. Current agreements allow satellites to be positioned every 2 degrees over the populous equatorial coverage areas.

The resulting proliferation in the number of equatorial satellites puts a significant burden on antenna designers. The 2- degree equatorial spacing requires that antennas have very narrow beams, in addition to low sidelobes, in order to prevent interference to transponders on adjacent satellites. The transmitters and receivers used in these satellites are very sophisticated, and must utilize adaptive techniques in order to maintain predetermined signal-to-noise ratios without interference problems.

The RF spectrum assigned to satellites includes band segments from 400 MHz to over 30 GHz. The most densely used portions are in the region between 1.5 and 12.5 GHz. Because of the limited availability of electric power, satellite transmitters tend to be limited to about 50 to 100 watts, but antenna gains of 20 dB or more are not uncommon. Obviously, satellites intended to provide broad geographic coverage cannot employ highly directive antennas. Some key applications for modern satellites include the following:

Earth resources and position location. Most of us have seen some of the dramatic photographs taken from earth resources satellites showing crops, earthquake faults, volcanoes, and of course, daily weather patterns. These satellites rely on a wide variety of sensors employing optical, thermal, and radar techniques to form images of the earth's surface. In most cases the resolution is sufficient to reveal major highways, airports, and other key land-based features.

Most of these satellites operate in the 137 MHz and 1.6 GHz portions of the spectrum. Fairly simple receiving equipment will permit the home experimenter to receive exciting weather photographs from the APT and GOES satellites.

An important additional function of this general class of satellites is mapping and position location. Virtually any satellite with high resolution sensors can function as an excellent mapping satellite since its orbit can be precisely determined and its position accurately known at any given time. Twenty continuous years of this process has resulted in highly accurate mapping of the earth's entire surface, permitting significant improvements in the efficiency and safety of commercial shipping and aviation. A series of Global Positioning Satellites (GPS) is being developed by the United States; when its 18 satellites are completed and in place a few years from now, the signals from these satellites will enable appropriately equipped vessels to determine their exact position on the earth's surface to within about 100 feet (30 meters).

Telecommunications. It is the telecommunications function performed by modern satellites with which most of us are familiar. Telephone communications, data transmission, video links, and broadcast television now unite the world in a massive satellite-linked network. Satellites designed for this purpose do not originate signals, but are simply repeaters in the sky. They typically consist of a number of channels, called "transponders," that repeat signals in a 24 or 36 MHz band.

ADVERTISER'S INDEX AND READER SERVICE NUMBERS

Listed below are the page number and reader service number for each company advertising in this issue. To get more information on their advertised products, use the bind-in card found elsewhere in this issue, select the correct reader service number from either the ad or this listing, check off the numbers, fill in your name and address, affix a postage stamp and return to us. We will promptly forward your request to the advertiser and your requested information should arrive shortly. If the card is missing, send all the pertinent information on a separate sheet of paper to: *ham radio* magazine, Attn: Reader Service, Greenville, NH 03048.

PRODUCT REVIEW/NEW PRODUCT

*Please contact this advertiser directly. *Limit 15 inquiries per request.*

Please use before June 30, 1985.

Each satellite can have as many as ten transponders. This means that a single satellite may handle as many as 50,000 phone calls simultaneously. The availability of these high quality RF links has dramatically improved intercontinental voice and data communications. These links can now reliably transmit a page of FAX data to another part of the world in just a few seconds. Satellites enable commercial broadcasters to bring us important events from virtually every corner of the earth as they occur; the societal ramifications of this news immediacy are still developing.

Defense. Not to be upstaged, the military establishments have also been busy in space. Since profitability, in the commercial sense, is not an issue for military users, some of the tasks which satellites are called upon to perform are indeed exotic. These include such things as very high-speed data links (1 Gbit/sec.) and very high resolution imagery for reconnaissance purposes. Charged coupled sensors, cooled with liquid helium to increase sensitivity and mounted at the focal plane of precision lenses, provide orbital images with stunning detail. The ability of radar to penetrate cloud coverage and provide high resolution images is particularly attractive to military users. In some cases, the orbital path of the satellite is used to synthesize an antenna of large aperture and provide a simulated very narrow beam. This technique, called "synthetic aperture radar" (SAR), is capable of revealing subtle details, not visible by any other means, hidden in the earth's surface. Techniques of this type are used by defense organizations to locate submarines concealed below the ocean's surface.

Next time you look at the evening sky, be sobered by the thought that at that very instant, there are probably more conversations being repeated through satellites than if all the world's Amateurs were on the air at the same time. The pros do it without QRM. Now *that's* what I call a band plan!

ham radio

JUNE 1985 / $2.50

ham radio magazine

rock-solid STABILITY and TUNABILITY

in this issue: high-stability receiver BFO • compact IF sweep generator • graphical selection of mixer frequencies • speed up your code • designing Yagis with the C-64 • inexpensive elevation indicator • automatic temperature control • wind your own transformers — inexpensively • plus W6SAI, W1JR, KØRYW, and the Guerri report

JUNE 1985
volume 18, number 6

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ham radio
magazine

contents

REFLECTIONS

our own miniseries

Anyone who watches TV has seen, I'm sure, at least one miniseries — be it *The Thorn Birds*, *Wallenberg*, or *SPACE*. Well, *ham radio* is pleased to bring you another. . . but with this one, happily, you won't have to wait until the next evening to see the continuation. We call it "sources for everyone."

Sources are popular items for Radio Amateurs. In fact, I defy you to show me a transmitter or transceiver that doesn't have at least one. In this issue we examine three different kinds of sources, each useful in its own way and each simple enough to be built by the "average" ham (whoever he may be).

Our first source — described in Al Helfrick's article, beginning on page 18 — answers the paradoxical question, What is rock-stable, yet varies in frequency? It's a voltage-controlled oscillator that uses a ceramic resonator for the frequency determining element. To say it's as stable as a crystal oscillator would be an exaggeration, but it does offer a significant improvement in stability over a conventional LC type device. Throw in the additional advantages of low microphonics, high *Q*, low cost in a small package, and you begin to get an idea of why both the author and I are interested in this device. Though it can be varied in frequency, its variation is limited by the series and parallel resonant frequencies of the resonator, which is about 7 percent. (Not bad considering the insignificant frequency pulling capability of a crystal oscillator.)

If we do need that high stability, Peter Bertini shows us how we can achieve it by using two crystal oscillators in a Colpitts configuration. The author, in designing BFO circuits, found an alternative to using expensive BFO crystals while still retaining the versatility of a variable BFO. In this application (a receiver BFO), very little frequency shift is needed, and it's nicely achieved by varying the voltage on the varactor elements. The difference frequency of the two crystals becomes the exact center of the variable frequency range needed. Tie that concept in with normal good oscillator design practices and we're left with a useful circuit that's inexpensive and quite stable.

The conclusion of this miniseries came out of a need for a simple and practical sweep generator that can be swept across a broad frequency range, yet still exhibit good linearity and constant output. From the Netherlands comes an article by Hans Evers, who discovered an old circuit — described perhaps for the first time — in a 1949 issue of *Wireless World*. The circuit, known as the Butler oscillator, is varied in frequency through control of an out-of-phase RF current that excites a (secondary) coil coupled to the main inductive element, thereby producing an effective turns cancellation. This current, in turn, is controlled through the unbalancing of a differential amplifier. All in all, an interesting technique.

If you like this miniseries, let us know we'll be glad to bring you others. What subjects would you like us to cover?

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

AMATEUR RADIO WILL FLY WITH THE SPACE SHUTTLE IN JULY after all, NASA has announced. However, astronaut Tony England, WØORE, will be limited to 2-meter operation only, as the installation of additional Amateur antennas for other bands in the shuttle's payload bay turned out to be too costly as well as a logistically difficult task.

Slow-Scan TV Is The Big Addition On This Shuttle Flight; it's even quite likely that SSTV transmission time will exceed that for voice, with scenes from inside the spacecraft to be sent during orbits when WØORE is busy with other tasks and can't be on the air for voice contacts. Both FM voice and SSTV will be via 2-meter hand-held radios, using the same window-mount antenna used by W5LFL in his pioneering operation.

Little If Any "General" 2-Way Operation Is Planned for this flight. Instead, WØORE's contact emphasis will be with school and club groups, to provide the maximum Amateur involvement. The ARRL will act as liasion to schedule such contacts; contact them in Newington for details. Pre-flight publicity and media contact during the flight will also be handled by the ARRL, but primarily through the League's Washington office.

NASA Expects WØORE To Be On For Between 10 And 20 Passes, and is currently working out his scheduling. A principal limitation will be access to the shuttle's overhead window, which won't be available at all for Amateur antenna use until the middle of the fourth day of the flight. At presstime launch was still scheduled for mid-July.

THE 24-MHZ BAND WILL BECOME AVAILABLE AND 30 METERS "OFFICIAL" JUNE 22, the FCC decided April 25. The new 24890-24990 kHz slot is for General and higher, with phone above 24930; power will be 1500 W PEP. 30 meters remains CW/RTTY only, General and above limited to 200 W PEP. No action was taken on 420 MHz changes or the new 902 MHz band.

International Broadcast Stations May Use 7.1-7.3 MHz In Region 3 (Pacific) areas it administers, the FCC decided April 4. In its Report and Order on Mass Media Docket 84-706 the Commission did specify, however, that broadcast stations operating under the new ruling must beam transmissions away from Region 2. The effective date was May 16.

Expanded 40-Meter Phone Privileges For Caribbean Area Amateurs are proposed in a new FCC Notice of Proposed Rulemaking. In PR Docket 85-104 Amateurs in U.S.-licensed areas of the Caribbean would have their phone band extended down to 7075 kHz; in its proposal the FCC is looking for Amateur input on how the additional frequencies should be utilized.

Comments On PR Docket 85-104 Are Due At The FCC June 17; Reply Comments July 17.

VOLUNTEER EXAMINER COORDINATOR PERFORMANCE MAY SOON be the subject of critical FCC review, now that the program has been on line for a year. Criteria to be used in the review will probably include activity, pass rates, adherence to the rules, integrity, and quality of the paperwork provided to Gettysburg. Some VECs are not expected to fare very well.

A Visit To Gettysburg For VEC Representatives is also being considered, probably for sometime in August. Such a visit would provide the VECs and FCC licensing people an excellent opportunity to discuss each other's problems and review procedures.

DeVry Is Actively Seeking VEs Throughout The U.S., following its accreditation as a national VEC. Interested VEs should call Jim Georgius, W9JUG, at DeVry between 12 noon and 7 PM Chicago time, at (312) 929-8500, ext. 251.

"AUTOMATIC REMOTE CONTROL" FOR AMATEUR OPERATIONS ABOVE 29.7 MHZ has been proposed by the FCC in PR Docket 85-105. Expanding on an ARRL petition that had sought automatic control for Amateurs using digital communications, the Commission is actually looking for Amateur input as to just how far the concept of automatic control should be extended.

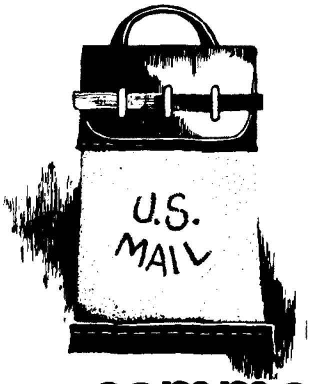
Comments On PR Docket 85-105 Are Due at the FCC June 25, and Reply Comments July 25.

THE COMMENT PERIOD ON FCC'S "NATIONAL FREQUENCY COORDINATOR" proposal is being extended at the request of the ARRL. The new dates had not yet been set at presstime.

AMATEUR RADIO'S SPACE PROGRAM WILL BE THE SUBJECT of the Teleconferencing Radio Net (TRN) June 14. Top AMSAT satellite specialists such as W3IWI, W6SP, WA2LQQ, W3XO, K8OCL, and others will participate in the comprehensive program, which will almost surely include the latest news on WØORE's forthcoming Space Shuttle operation.

Over 200 Repeaters Across North America Now Provide TRN programs to their users; any repeater group wishing to join the TRN should write Timothy Lowenstein, WAØIVW, TRN Net Manager, c/o Midway ARC, Box 1231, Kearney, Nebraska 68847-1231.

DAYTON HAMVENTION'S "AMATEUR OF THE YEAR" IS W8ACE/4, John Willig, who's being recognized both for his role as "father" of the Hamvention back in 1951, and for his continuing efforts in maintaining the "Dayton Net" that meets three times weekly to keep former Daytonians in touch with each other. The Hamvention's Special Achievement Award went to Judy Frye, KG8P, for her leadership role setting up DARA's VEC program; the Technical Excellence Award was presented to Rich Whiting, WØTN, for having developed the Teleconferencing Radio Net. Congratulations to all for well-deserved honors!



comments

How can Amateur Radio be improved? Here's what some readers suggested in response to questions raised by K2RR and N1ACH in "Reflections" (February and March, 1985). **KA1LBO**

I'd like to see us stop all the concern for increasing *numbers* of hams and instead, try to improve the quality of operators we already have.

Vern A. Weiss, WA9VLK
Kankakee, *Illinois*

. . . I would like to see the spirit of Amateur Radio as it was in the 1930's restored: *service*, experimentation, operating skill, courtesy, and good fellowship.

I. L. McNally, K6WX
Sun City, California

. . . There seems to be an age of curiosity — maybe between 10 and 15 — when a youngster is thirsty to drink up the experience of the adult world. At that age the young person is curious about the entire world. We, the adults, can "turn that kid on" to electronics, computers, cameras, art, or music — or they can turn on to drugs. It's up to us. . . . My point is that we must make a solid effort to get into the elementary schools, into the summer camps, into the neighborhood community centers, to "turn on" the kids of America to the fun of our hobby. This will, in turn, capture the talents of our youth *and reignite the spirit that* put a man on the moon!

Marvin Feldman, WG4Q
Annandale, Virginia

In days past, there was a "romance of the airwaves" — the excitement of doing what ordinary people could not do — *and you could do it*, if you knew enough and worked hard enough.

We need to find a new romance, a new excitement.

John Telford, W3TJ
Swampscott, Massachusetts

Here's what I think is important right now — more important than technical development, if we have to choose: *we must learn how to lobby*. We hams must learn individual and group lobbying techniques. Lobbying is power, and groups have more power than individuals.

[We have to] get our emergency nets working and then let the public know what we're doing for them. When the public needs communications for various activities, we should *overwhelm* them with support, and then *publicize* our contribution.

Ken Uthus, KT7E
Nine-mile Falls, Washington

The way I entered ham radio was via SWL activities. I was in junior high school and could not afford anything but a receiver. I still listen to SW on a day-to-day basis and enjoy it. Perhaps emphasizing such activities will stimulate interest. . . .

Thomas M. Hart, AD1B
Dedham, Massachusetts

. . . What Amateur Radio needs is a new frontier. . . . Why not put the total emphasis on space [communications]? Maybe radio telescope operation. Let's restore the adventure of Amateur Radio to attract the new breed of high-tech communicators.

Fred N. Ackerman, W3JHN
Silver Spring, Maryland

I've been a ham for 54 years. . . . Ten years ago I got the SSTV bug and the biggest thrill from ham radio, ever. What happened to SSTV articles in most of the magazines?

Francis M. Duffy, AB3J
Johnstown, Pennsylvania

My main operating interest is CW. I perceive considerable anti-CW pressure these days . . . more publicity and development for very narrow band synchronous CW would help . . . also improvement in the political climate for this mode. . . .

To my mind, hand-sent Morse is more basic than BASIC.

Bruce Boyd, W3QA
Ellicott City, Maryland

Let's stop recruiting non-technical people into our fraternity. . . . Let's try to appeal to an average level of technical expertise — on upward.

Mike Kitsko, K6VGQ
Cerritos, California

A link is needed between HF, DX, 2 meters, CB, SSTV, etc. A geostationary satellite could do *just that*.

Greg Waits, WB6EPE
Anaheim, California

The number of licensed Amateur Radio operators may be declining statistically but I feel sure that there are many others out there just as enthusiastically involved as we are here. The new VEC program will reverse the [negative] trend. . . . Amateur Radio clubs all over the country can and will take advantage of the VEC program to promote and assist with the licensing of new Amateur Radio enthusiasts.

Let's not be too concerned with declining numbers. It's only temporary. We, the members of the World Amateur Radio Fraternity, will see to that.

Bob Ruedisueli, W4OWA
Vienna, Virginia

. . . You may want to consider that the new 10-year license (term) will skew the figures.

John D. Gallivan III, N4DGS
Fairfax, Virginia

. . . What's wrong with Amateur Radio? Not a damn thing . . . it's just *different*.

Michael Vuksich, WØVEV
Duluth, Minnesota

voltage controlled oscillator uses ceramic resonators

Try these old favorites in tuned circuits

Since the earliest days of radio and electronics the LC tuned oscillator was the circuit of choice whenever a wide tuning range was needed. When stability was paramount, the crystal oscillator reigned supreme.

Over the years the frequency controlling element has changed; oscillators controlled by cavities, surface acoustic wave (SAW) devices, YIG (*Y*ttrium *I*ndium *G*arnet) and other garnets, and a host of other specialized devices have been developed. Very often a system design would benefit from an oscillator that had the tuning range of an LC oscillator and the stability of a crystal oscillator.

The phase-locked synthesizer is a system in which the accuracy and stability of a quartz crystal is transferred to a wide frequency range oscillator, permitting a large number of accurate and stable frequencies to be generated. Even in this application, the flaws of an oscillator cannot be totally eliminated. If the oscillator suffers from phase noise, microphonics, or other spurious outputs, some vestiges of these problems will remain.

The voltage controlled oscillator can be constructed from any type of oscillator, but is typically an LC oscillator with a varactor diode as part or all of the tuning capacitance. This gives the broadest tuning range but seldom exceeds a ratio of 2:1. At the other end of the spectrum (minimum tuning range), a crystal oscillator with varactor tuning provides excellent stability, very low phase noise, and freedom from microphonics, but has very limited tuning range. A rule of thumb for a variable crystal oscillator is a frequency variation of no more than 0.1 percent total. If the frequency were limited to 20 MHz (this is approximately the limit for fundamental mode crystals), the total variation would be 20 kHz. For higher frequencies, overtone crystals must be used and the ability to pull the frequency of an overtone crystal is considerably less than that of a fundamental mode crystal.

ceramic oscillator provides compromise

Somewhere between the crystal oscillator (which can hardly be moved at all in frequency) and the LC oscillator (which has a broad variation in frequency but suffers from phase noise and microphonics) is the ceramic resonator oscillator. The ceramic resonator is not a new device; for many years ceramic resonators have been the heart of the ceramic filter, which is primarily used for wideband FM IF systems found in entertainment radio and television receivers. The ceramic resonator as a single tuned circuit is finding applications in timing circuits, replacing more expen-

By Albert D. Helfrick, K2BLA, R.D. 1, Box 87, Boonton, New Jersey 07005

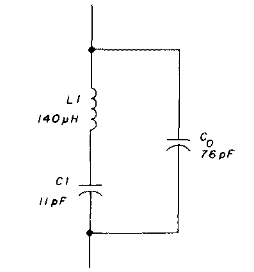


fig. 1. Equivalent circuit of a 4-MHz ceramic resonator.

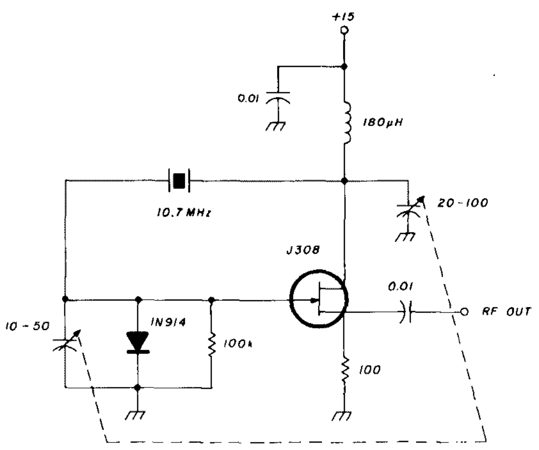


fig. 2. Mechanically-tuned Pierce oscillator.

sive quartz crystals in those applications in which the extreme precision and stability of the quartz resonator is not required.

Although the ceramic resonator is similar to a quartz crystal, there are several important differences. The equivalent circuit of the ceramic resonator shown in **fig. 1** is identical to the quartz crystal except that the parameter values are somewhat different. The series resistance is on the order of 6 ohms for a 4-MHz resonator, which compares to a quartz crystal's 15 ohms. This lower value at first appears to indicate a higher Q than with the quartz crystal if it weren't for the value of the equivalent inductance, which is typically 140 microhenries. In a quartz resonator at a frequency of 4 MHz, the equivalent inductance could easily be on the order of 1 henry, which is some 7000 times higher. This results in the quartz crystal having a Q of as high as 500,000. In the ceramic resonator a series capacitance of 11 pF provides the series resonant frequency while a 76-pF shunt capacitance, plus any circuit capacitance provides the parallel resonant frequency. The Q of the equivalent circuit is on the order of 600.

advantage of the ceramic resonator over the LC oscillator

There are two inherent advantages in using a ceramic resonator rather than a conventional LC tuned circuit. First, it is nearly impossible to build inductors in the 140 µH region that are small in size and exhibit Qs of 600. Typically a 140-µH inductor has a Q of one-tenth that of a ceramic resonator. Secondly, a 140-µH inductor, in order to be of reasonable size, would use a tightly coupled ferrite core. The core would be attached to a shielded can and fitted with an adjuster. This would translate mechanical movement of the core into changes in the inductance and distributed capacitance. An oscillator constructed with this type inductance would tend to be microphonic.

There are broad variations in the parameters of ceramic resonators, especially when resonators of different frequencies are compared. However, ceramic resonator Qs will generally range from 500 to 5000, values not achieved using conventional LC tuned circuits.

The significant disadvantage of the ceramic resonator as an oscillator tuned circuit is the ability of the resonator to be varied in frequency. Ceramic resonators are typically operated in parallel resonance, which is characterized by the resonant frequency of the equivalent inductance resonating with the two equivalent capacitances, C_0 and C_1. The 76 pF capacitance shown in **fig. 1** is internal to the resonator and cannot be altered in any way. Consider a situation in which a varactor is placed across the resonator to vary the frequency. The highest possible resonant frequency (f_p) when operating the resonator as a parallel resonant circuit occurs when there is *no* external capacitance. This produces a resonant frequency of:

$$f_p = \frac{1}{2\pi\sqrt{\dfrac{L_1C_0C_1}{C_0 + C_1}}} \tag{1}$$

There is a practical limit to how much external capacitance can be placed across the resonator, but for the purpose of gaining insight into how far the resonator can be theoretically pulled, assume the external capacitance can be as high as infinity. With infinite external capacitance, the parallel resonant frequency approaches the series resonant frequency (f_s) which is:

$$f_s = \frac{1}{2\pi\sqrt{L_1C_1}} \tag{2}$$

determining range of operation

If the ratio of the highest frequency to the lowest frequency were taken, the result would be:

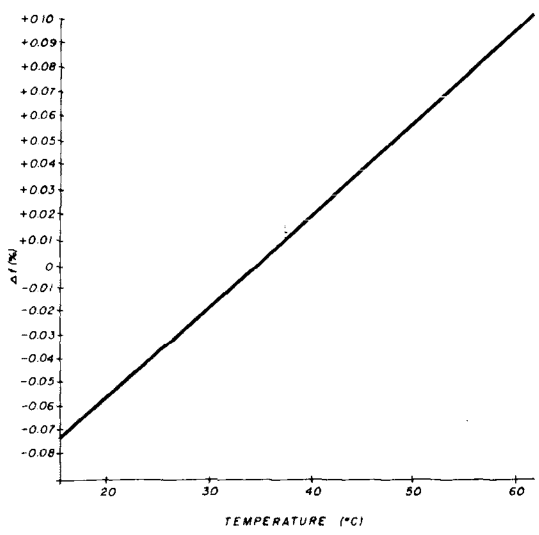


fig. 3. Temperature-induced frequency variation of mechanically-tuned resonator oscillator.

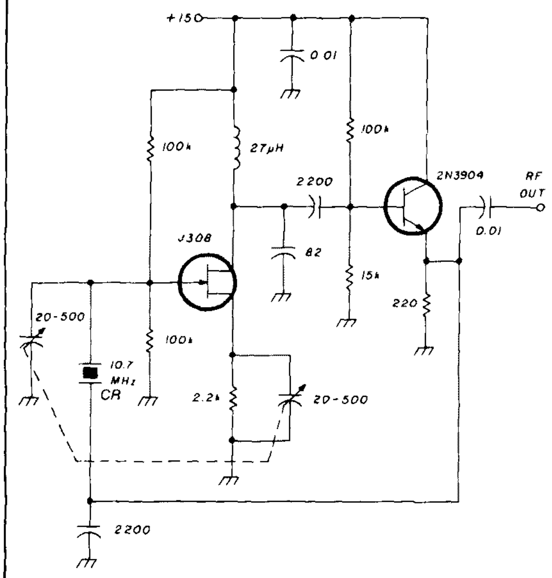


fig. 4. Mechanically-tuned ceramic resonator oscillator.

$$\frac{f_p}{f_s} = \sqrt{\frac{C_0 + C_1}{C_0}} \qquad \textbf{(3)}$$

Using the typical values shown in **fig. 1**, the maximum frequency range (and consequently change from nominal frequency) is about 7 percent total for a 4-MHz resonator. Compared to the quartz crystal, this is a very large variation but rather insignificant to the octave or more variation available from an LC oscillator.

The previous discussion determined the maximum theoretical variation. However, it is not possible to apply an external capacitance from zero to infinity. The lower value of capacitance is limited by the circuit capacitance and the *residual* capacitance of the varactor. This discussion examines varactor tuned oscillators, but there is residual capacitance in mechanical capacitors as well. In addition, there are practical limits to the upper bound of the external capacitance and from empirical data this can be safely taken to be three times the equivalent shunt capacitance of the resonator. When a large amount of external capacitance is used, the dissipation factor of the external capacitor becomes a critical parameter in the amount of energy that can be transferred into the ceramic resonator. As a design example, assume the minimum external capacitance is 30 pF and three times the internal 76 pF or 228 pF is the maximum capacitance that is allowed. This would result from a varactor diode variation of about 200 pF. If the residual circuit capacitance is 10 pF exclusive of the varactor capacitance, the varactor diode capacitance must vary from 20 pF to 218 pF. This requires a tuning ratio of 10, which is attainable from a hyper-abrupt junction diode. The typical frequency variation would be about 3 percent or approximately 120 kHz at 4 MHz.

comparing actual circuits

To test the theory and to assess the ability of the ceramic resonator to serve as the frequency controlling element of an oscillator, several conventional capacitor-tuned oscillators were constructed. One example is shown in **fig. 2**. The ganged variable capacitor tunes both the gate and drain of the FET oscillator and thus maintains a constant feedback ratio. This oscillator produced a tuning range of 190 kHz from 10.7 MHz, the specified resonant frequency of the oscillator, to 10.89 MHz. This simple circuit, also illustrated a design limitation of the ceramic resonator — frequency drift with temperature.

Ceramic resonators are frequency-stabilized over temperature by using special −4400 ppm/degree C ceramic capacitors as the feedback elements. Although the temperature characteristic of the mechanical capacitor was not known, it did *not* correct the frequency drift and the results are shown in **fig. 3**. In spite of its frequency dependence on temperature, the oscillator had some important characteristics. The oscillator had a low level of phase noise and was relatively immune from microphonics. It is these two characteristics that make the oscillator attractive as a VCO for a phase locked loop frequency synthesizer where the temperature dependence can be eliminated.

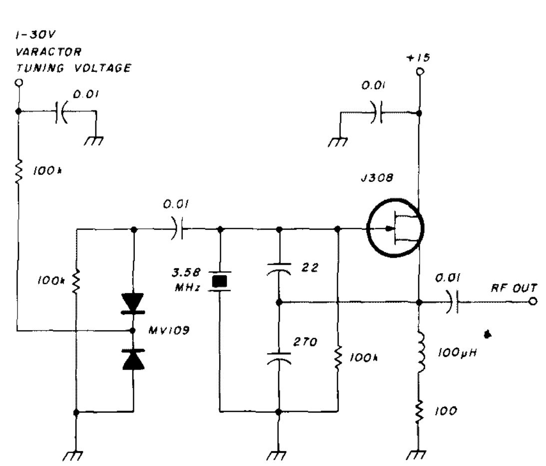


fig. 5. Colpitts oscillator using varactor tuning and a 3.58-MHz resonator.

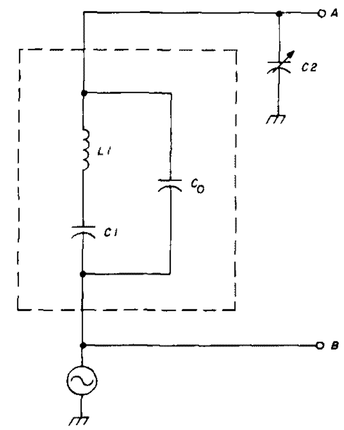


fig. 6. Ceramic resonator circuit with AC voltage source and capacitive load.

A second mechanically-tuned oscillator was constructed and is shown in **fig. 4**. One of the problems associated with an oscillator using a ceramic resonator and, to a certain extent, a crystal resonator, is that when a large amount of external capacitance is added the amount of feedback is reduced and oscillation can become unsteady (stops oscillating). The oscillator shown in **fig. 4** represents an attempt to alleviate this problem. A dual 500-pF capacitor was used as the tuning element across a 10.7-MHz resonator. The gain of the FET amplifier was varied by increasing the capacitance from the source to ground using half of the dual capacitor. As the capacitance across the resonator is increased, the gain of the amplifier also increases and the oscillations are stable over the entire 500 pF range of the capacitor. The tuning range of this oscillator was 325 kHz at 10.7 MHz.

A varactor-tuned Colpitts oscillator as shown in **fig. 5** was constructed to evaluate the ability of the ceramic resonator to serve as a VCO. A 4-MHz resonator was tested, and the results agreed substantially with theory. Using a pair of hyper-abrupt varactor diodes, it was possible to obtain nearly a 100 kHz frequency shift, which agrees with the theoretical calculation of 120 kHz. With a large variation in capacitance an unsteady oscillatory condition was evident. This is because the feedback capacitors were much smaller than the maximum capacitance of the varactors. It is mandatory that the fixed capacitances across the ceramic resonator be kept to an absolute minimum.

In the development of an oscillator that provides a maximum frequency shift using a ceramic resonator as the tuned circuit, consider the circuit shown in **fig. 6**. The equivalent circuit of a ceramic resonator is

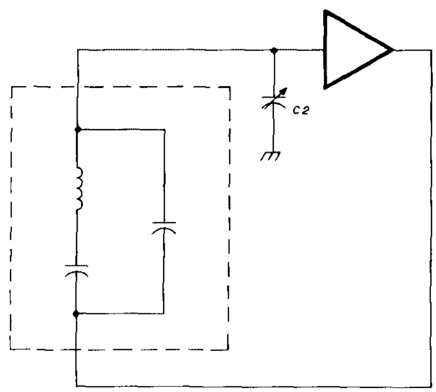


fig. 7. Ceramic resonator oscillator.

shown in series with an AC voltage source and an external capacitance which represents the varactor. Since the impedance of the voltage source is zero, the external varactor capacitance is essentially in parallel with the internal capacitance, C_0. The voltage at point *A* is exactly out of phase with the voltage at point *B* at the parallel resonant frequency of the inductor, L_1, and $C_0 + C_2$ in series with C_1. Only one resonant circuit is shown in **fig. 6**. The usual series resonant circuit of L_1 and C_1 is not a factor here. If there were a resistance between point *A* to ground, in parallel with C_3, there would be a second point at which the phase of the voltage at *A* would be the same as the voltage at point *B*, and that would occur at the series resonant frequency determined by L_1, and C_1.

Figure 7 shows the equivalent circuit of the ceramic

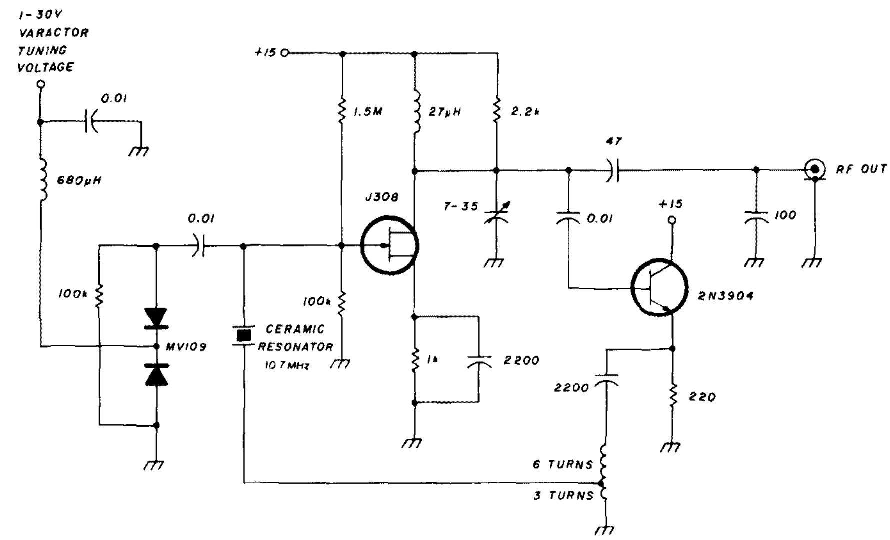


fig. 8. Varactor-tuned 10-MHz ceramic resonator oscillator.

resonator except that the AC voltage source has been replaced with the output of an inverting amplifier and point *A* from the equivalent circuit of **fig. 6** has been connected to the input of the inverting amplifier. To satisfy the Barkhausen criteria for sustained oscillations, the total phase shift around the loop should be 0 degrees, which will occur at the parallel resonant frequency of **fig. 6**. In addition, the total loop gain should be zero dB, which can be satisfied by the buildup of self bias, which automatically reduces the circuit gain, as in the case of any classical oscillator. For C2 to have the greatest effect on the oscillator frequency, the input capacitance of the amplifier must be as small as possible and the output impedance of the amplifier as low as possible.

The oscillator shown in **fig. 8** satisfies the criteria for a low output impedance and small input capacitance. The FET input amplifier has fixed bias with source feedback. This provides a very high input impedance with very low capacitance. The FET amplifier drives an emitter follower which, in spite of the fact that it has a low output impedance, feeds a transformer with a 3:1 turns ratio for a nine-fold impedance reduction. The result is an impedance at the ceramic resonator of a few ohms maximum.

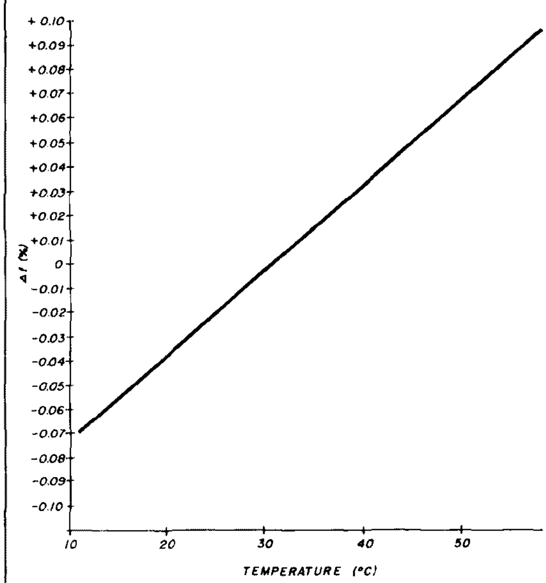


fig. 9. Temperature dependence of frequency for the varactor-tuned oscillator shown in *fig. 8*.

The varactor-tuned ceramic resonator oscillator has a significant frequency-temperature coefficient, as would be expected in light of the results of fig. 3, and is shown in **fig. 9**. The tuning range of the VCO is approximately 232 kHz, with a temperature coefficient of 350 Hz per degree centigrade. When using this circuit as a VCO, the entire 232 kHz range cannot be used because some of the tuning range must be sacrificed for the temperature dependence. If the required tuning range were 200 kHz, leaving 32 kHz for temperature variation, the resulting temperature variation would be more than 90 degrees C, which is sufficient for any Amateur application.

It may appear that a temperature compensating ceramic capacitor could be used for the reduction of

temperature dependence. This is done for single-frequency oscillators, but the use of a shunt capacitance for temperature compensation will reduce the amount of frequency variation possible. In an oscillator with a large tuning range, temperature compensation must be achieved with a phased locked loop or other external means of frequency stabilization.

To what advantage is the ceramic resonator oscillator when it has a significant frequency-temperature coefficient and has a tuning range of only a few percent? First the phase noise of the ceramic resonator oscillator is excellent, as shown by the spectrum analyzer photograph in **fig. 10**. It would be nearly impossible to create a spectrum this clean with a conventional LC tuned oscillator. In addition, the freedom from microphonics is excellent. As a test, the fourth harmonic of the 10 MHz VCO shown in **fig. 8** was received with a VHF/FM monitor receiver. This receiver is designed to function with 5-kHz peak frequency deviation, and because the fourth harmonic is being tuned a 1.25-kHz peak deviation will produce the full audio output in the receiver. The oscillator was tapped with a pencil and absolutely no audio was heard from the receiver. This is certainly not a scientific test, but to gain some insight into how this compares to a conventional LC oscillator, the printed circuit board in the monitor receiver was tapped and produced a loud clang in the receiver. In fact, if the volume of the monitor receiver were advanced to nearly full scale, a steady howl would emanate from the speaker because of acoustic feedback from the speaker to the local oscillator. Although the test was not scientific, it is clear that the ceramic resonator oscillator is quite free from microphonics.

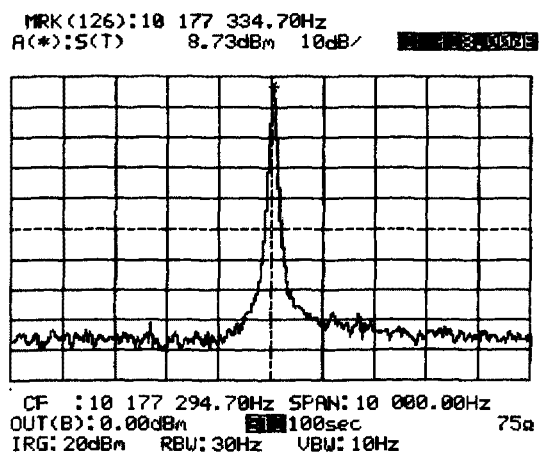


fig. 10. Spectrum analyzer photograph of a ceramic resonator VCO without a phase locked loop. The VCO's actual performance is difficult to determine due to the noise floor and resolution limitations of the instrument.

applications

There are many applications for a narrow band low-noise oscillator. An obvious choice is to provide a VCO for a narrow range synthesizer such as a single Amateur-band frequency source. The new 10 MHz band is an obvious choice, since the VCO shown in this article is directly applicable. Another application is shown in **fig. 11**. Here the ceramic resonator oscillator is used in a frequency synthesizer with 100 Hz resolution. The 10-MHz range of the ceramic resonator is translated to the 6-meter Amateur band by heterodyning it with a 40-54.6 MHz phase locked frequency synthesizer. Because of the rather high (100

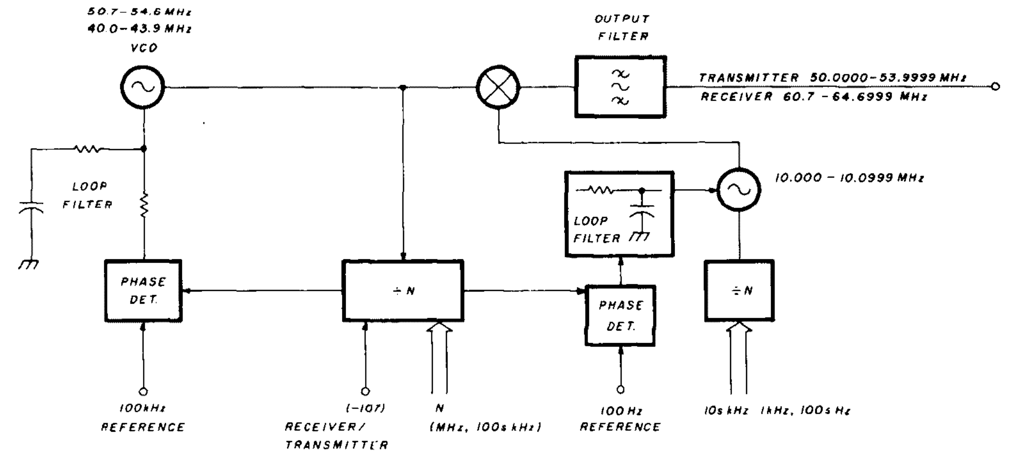


fig. 11. Ceramic resonator oscillator used in a 100 Hz resolution 50-MHz synthesizer.

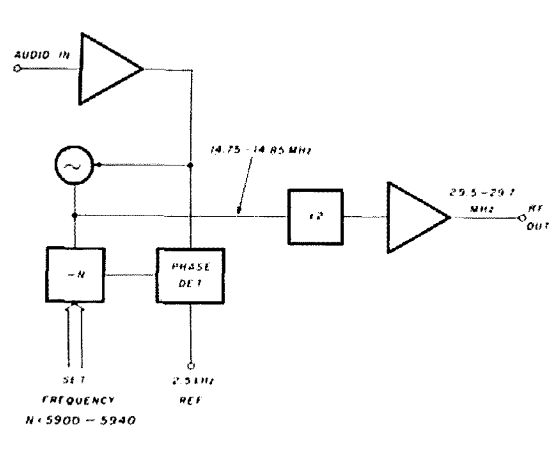


fig. 12. A 20-meter FM transmitter using a ceramic resonator oscillator.

kHz) reference frequency used for the higher frequency synthesizer, the microphonics and phase noise from that synthesizer can be reduced to negligible amounts. The relatively low frequency of the 100 Hz reference of the second phase locked loop does not allow for the loop to remove any microphonics of the inherent low-microphonic VCO using the ceramic resonator. Any synthesizer using a low frequency reference can benefit from the ceramic resonator oscillator.

A second application is shown in **fig. 12**. This example shows a 10-meter FM transmitter using a ceramic resonator. The basic 15-MHz frequency of the ceramic resonator oscillator is doubled to 29.5-29.7 MHz and synthesized using a 2.5 kHz reference. As in any FM transmitter where the phase locked loop is modulated directly, the loop bandwidth is restricted to a frequency below the audio frequency range, which opens up possibilities for microphonics and phase noise. This is demonstrated in most synthesized FM VHF equipment by the fact that despite putting in epoxy or other compound, the VCOs are often microphonic and noisy.

The ceramic resonator oscillator has proved to be an interesting device for several applications and it is suspected that other applications will appear. No attempt was made to temperature-compensate the basic oscillator circuit. It is possible that a thermistor or other temperature sensing device could be used in conjunction with the varactor tuning voltage to reduce the temperature dependence of the ceramic resonator to an acceptable amount. In addition to its use in an oscillator circuit, the ceramic resonator has important applications in any circuit in which a conventional LC tuned circuit would be used.

ham radio

a high-stability BFO for receiver applications

Two VCXOs work together to provide high performance

The lower frequency IF (under 500 kHz) remains popular for many receiver applications, generally as the last IF of a single or multi-conversion scheme.

A typical Amateur application will usually require two or three discrete BFO frequencies — one for each sideband and one for CW. Good mechanical filters, either imports or surplus models made by Collins-Rockwell, are readily available. Unfortunately, while good filters are available, their companion BFO crystals are seldom offered; custom-made BFO crystals for under 500 kHz can be obtained, but only at a premium price.

Many hams have avoided the high cost of crystals by designing variable BFOs to take their place. While this is a reasonable alternative, there are disadvantages. If, for example, the filters will do double duty in the transmitter portion of a transceiver, BFO crystals would have to be used in order to ensure predictable filter performance. Most filter manufacturers specify the filter 20-dB attenuation points as the recommended BFO frequencies for SSB operation. Shifts in the BFO frequency could cause loss of carrier suppression or an undesired audio bandpass. A synthesized BFO circuit for 9-MHz IFs was described in *ham radio* several years ago, but it did not have provisions for variable tuning.[1] For certain receiver applications — especially for radioteletype (RTTY) or serious CW work — a variable BFO is a desirable feature. But we are still at the mercy of the long and short-term drift characteristics of a free-running LC oscillator. In light of the considerable investment a homebrewer makes for a set of decent filters, it would be false economy to compromise an otherwise good receiver by using a second-rate BFO design.

alternative approach — two 15-MHz oscillators

I've been building a general coverage receiver for about five years — it's one of those low-priority projects that just sits on the back burner and is worked on only during periods of extreme ambition. During my last brain-storming session I tackled its BFO circuits. My particular application required a variable frequency source between 5.593-5.597 MHz (for passband tuning) and the ability to preset for sideband generation in a planned companion exciter that would make use of the receiver's BFO, VFO, and HFO signals.

In designing the BFO circuits, I found an alternative to using expensive BFO crystals that would retain the versatility of a variable BFO. While my circuit uses two oscillators in the 15-MHz range, the particular frequencies are not of primary importance. (It would be wise, however, to avoid frequencies that fall on other receiver IFs or in the main tuning ranges.) Plated crystals in the 10 to 22-MHz range (fundamental frequency) will work the best; surplus crystals using pressure-plate mounts are not recommended.

What is important is that the frequency difference of the two crystals is in the exact center of the variable-frequency range we desire. I ordered two CS-1 grade crystals from International Crystal Manufacturing Co.,* and they seemed to be quite *willing* to *match* the error of the two crystals during production. This BFO system is readily adaptable to other popular IF filters in the 5, 8, or 9-MHz ranges by simply inserting the proper crystals into the circuit.

Both crystals are used in identical Colpitts circuits (fig. 1), except in that the frequencies are varactor controllable to a small degree. The total BFO shift is at least 4 kHz using the fundamental mode. This shift can be increased to about 6 kHz by installing 1.0 μH inductors (molded chokes) in series with the crystals. A circuit for overtone crystals in the 40-50 MHz range is shown, (fig. 2). The maximum frequency excursion is considerably lower — typically 1.5 kHz. Series inductance can be used to increase the tuning range of

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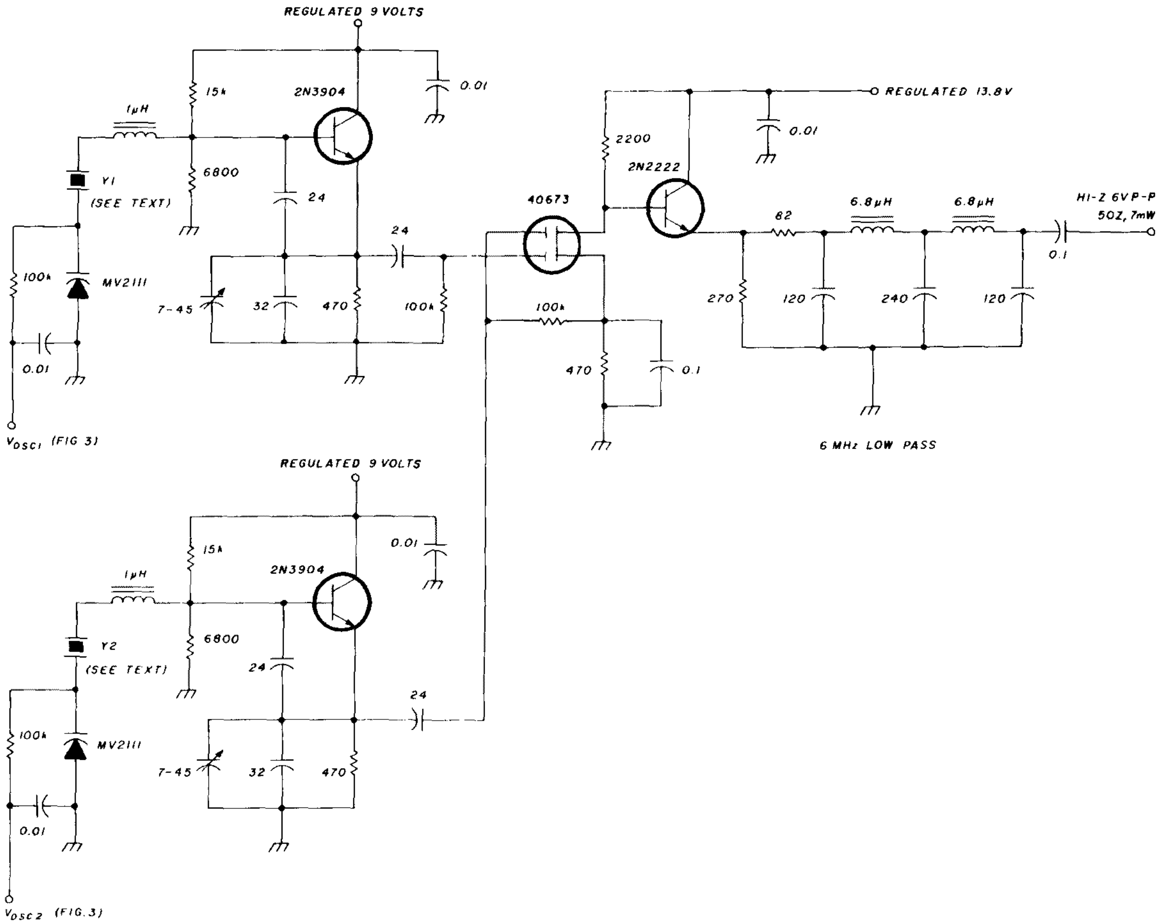


fig. 1. Complete VCXO schematic. Shown are the fundamental mode oscillators, the mixer stage, and buffer. As described in the text, crystals Y1 and Y2 are in the 10 to 22 MHz range. International Crystal type CS-1 3rd overtone crystals (ICM No. 471360) were operated on their fundamental mode in this circuit.

the overtone mode slightly, but the inductors should not be greater than 0.5 μH. It is likely that most BFO requirements can be met with approximately a 1.5-kHz tuning range.

With both varicap control inputs tied to 1/2 V_{CC} the BFO output should be at the exact difference frequency. If a small error exists it can be trimmed out by careful adjustment of the 45-pF trimmers in the Colpitts feedback network. Initially both trimmers should be set at midpoint, and any required trimming accomplished by adjusting both trimmers equally in opposite directions. NPO capacitors should be used in the oscillators. The oscillators should be powered from a 9 to 10 volt regulated source. Either a zener or three-terminal regulator will serve here.

*International Crystal Manufacturing Co., 10 North Lee, Oklahoma City, Oklahoma 73102.

SSB requires few kHz frequency change

Some common VCXOs (Variable Crystal Oscillators) use both variable capacitive and inductive elements to achieve a wider pulling range. Typically these circuits become somewhat unstable at the tuning extremes. In this BFO design, we require only a few kHz tuning range to meet most SSB BFO requirements. In my circuit two oscillators are pulled in opposing directions, effectively doubling the range normally expected from a single oscillator. Producing the two varactor tuning voltages could be accomplished using a dual-ganged potentiometer, but I chose the more elegant approach of using an IC voltage inverter to drive one of the oscillator varicap diodes. This permits the use of a single-stage variable potentiometer to control both oscillators. At first glance, the circuitry used

to accomplish this may appear to be more involved than necessary. This will be explained in greater detail further into this article (see **fig. 3**).

The mixer is a 40673 dual-gate MOSFET. An emitter-follower buffer stage provides about +7 dBm (5 mW) when driving a 50-ohm load; or about 6 volts P-P driving a high-Z termination. The low-pass filter shown is optimum for outputs in the 5-6 MHz range. These values can be scaled for other IF ranges. Suitable filters can be found in tables appearing in handbooks or electronic data books.

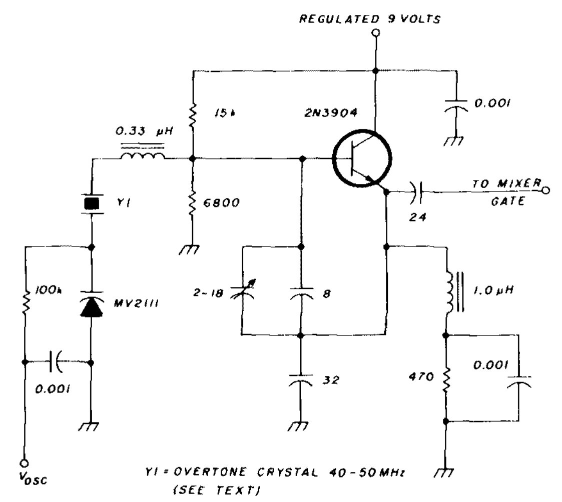


fig. 2. Optional overtone oscillator circuit for use in the BFO. Crystals again are ICM CS-1 (No. 471360) 3rd overtone cuts in the 40 to 50-MHz range. Using this overtone circuit, the maximum BFO tuning range will be limited to about 2 kHz.

good performance at low cost

Although not shown in the schematics, it would be simple to switch trimpots into the circuit in place of the tuning potentiometer to permit generation of fixed discrete BFO frequencies. Thus, one could switch automatically to a preset BFO frequency during transmit and have either preset or manual control of the BFO during receive. The real beauty of this circuit is in the long and short-term stability it ensures. Because of aging or temperature cycling, crystal or component variations should result in a nearly identical positive or negative frequency shift in both oscillators, but the desired difference-frequency will be closely maintained. Control voltages should be taken from a regulated low-noise source for best results. The noise-floor of this BFO will be better than most free-running designs, contributing to better post-filter IF noise performance.

The cost of this circuit compares favorably with the cost of individual low-frequency crystals. While the unit shown was built for 5.595 MHz, the basic scheme is useful for other HF and MF frequencies. The circuit is also useful for a variable frequency source in other

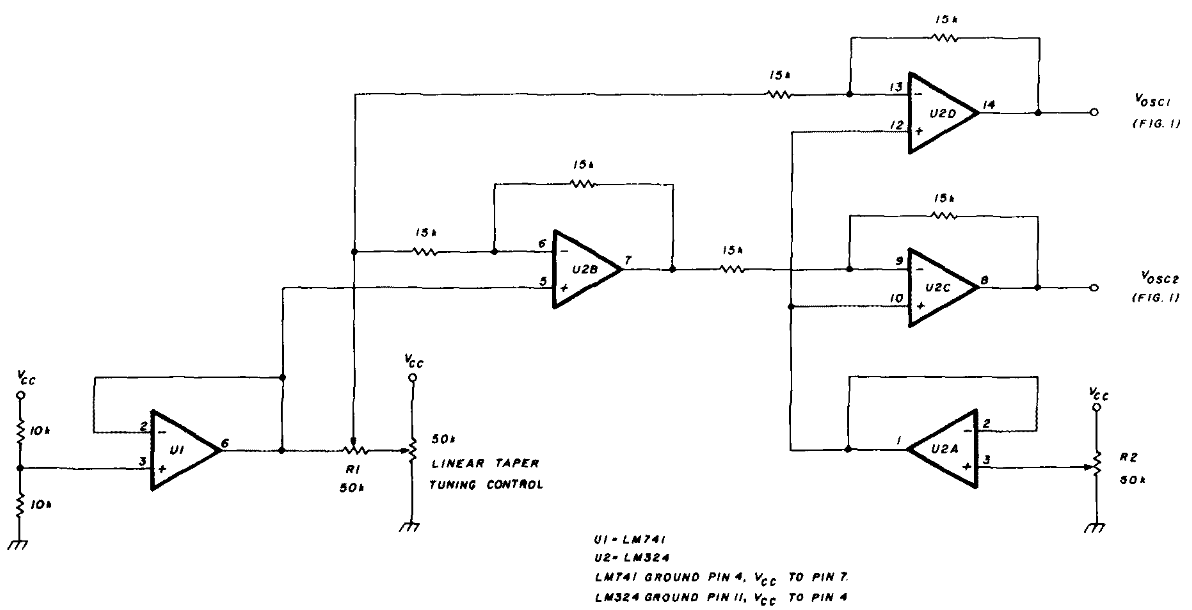


fig. 3. Control tuning voltages are provided by this circuit. The settings for R1 and R2 are discussed in the text. Shielded leads should be used to carry the voltage-oscillator signals to prevent incidental FM modulation from external AC field.

applications requiring good stability over a small frequency range, such as LOs for IF-variable passband or bandwidth tuners, or perhaps as a fine frequency-shifter in some PLL designs.

initial adjustments

The varicap diodes do not exhibit a linear change in capacitance for a given change in voltage over the entire available VCXO tuning range. Some empirical "cut-and-try" adjustments will be needed to obtain the desired results. The greatest change in varicap capacity occurs in the first few volts of tuning bias voltages; this is also the area of greatest non-linearity. Higher tuning voltages will result in smaller tuning ranges, while the tuning linearity will improve. Trimpot R1 sets the DC gain of the op amps (to control the maximum shift of the varicap tuning voltage). The op-amps should be powered from a well-regulated 15 to 24 volt supply. Maximum tuning range will be obtained with the higher voltage.

Trimpot R2 sets the DC offset of the two op amps used to drive the varicaps (to set the DC voltage at which the tuning voltage will start and end). The LM324 op amps were used because of their ability to reach near the power supply bus voltages (V_{CC} and ground) extending the tuning voltage range to those limits.

Trimpot R1 is adjusted to set the desired voltage swing for the varicap tuning produced by the full rotation of the tuning potentiometer. Trimpot R2 is used to set the bias point of the varicaps (the voltage midpoint between the tuning-voltage extremes). Regardless of the setting of R2, the center-tuning voltage for the two varicaps will be the same. For example, assume R1 was set for a 2-volt range, and R2 was set for a center-tuning voltage of 4 volts. The result would be that the noninverted varicap would swing from 3 to 5 volts, while the inverting varicap tuning voltage conversely would swing from 5 down to 3 volts, for a full rotation of the tuning potentiometer. If R2 were set for 7 volts tuning-center, the varicap tuning voltages would be from 6 to 8 volts and 8 to 6 volts, respectively. At first this may appear to be trivial, but remember that the tuning curves for the varicaps is not a linear function. This circuit ensures that any nonlinearity of tuning — on either side of center — will be the same, or symmetrical to each other. Thus, the related panel markings would be symmetrical and more pleasing to the eye, and the "tuning-feel" more natural.

reference

1. Raymond C. Petit, W7GHM, "Phase-Locked 9-MHz BFO," *ham radio*, November, 1978, page 49.

ham radio

a compact
IF sweep generator

A stable frequency source with excellent linearity and constant output level

In a **sweep generator** for the lower IF and HF bands, tuning diodes, or "varicaps" generally don't produce sufficient frequency swing with acceptable linearity. On the other hand, other methods of frequency sweeping — for example, frequency conversion, reactance circuits, and mechanical tuning — may cause the cost and complexity of a home-brew project to increase to the point of being both unaffordable and impractical.

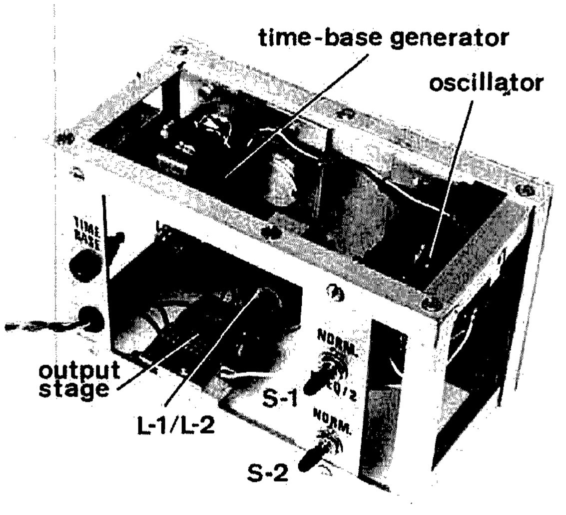
That is why, after coming across an old and half-forgotten oscillator circuit, I was pleasantly surprised to find that a large frequency sweep with good linearity isn't that difficult to build, as this simple little test instrument shows.

To keep the sweep generator as practical and uncomplicated as possible, I decided to use only one frequency band (100 to 200 kHz), using the harmonics of the basic signal for higher frequencies. This approach offers the advantage of using only one coil, without any switching circuit. It also permits the use of a simplified power output stage, which requires only an ordinary low-power transistor.

The principle of using harmonics works out very well. Because the frequency of a measured filter is generally known, confusion about the correct frequency is unlikely. In addition, the waveform of the output signal cannot alter the response of the filter under investigation. Interference from other than the wanted harmonic is not possible either, because the next harmonic is always at least 100 kHz away.

By making the sweep generator deliver pulses rather than sine waves, another advantage comes to light: it's easy to keep the output level of the different harmonics constant, regardless of frequency variations, by simply maintaining the wave form (duty cycle) of the pulse signal.

The little frequency sweeper has already demonstrated its value — not only by aligning IF strips in my equipment, but also by inspiring me to carry out several interesting experiments with crystal filters that would otherwise be very difficult to do.



Inside view of the IF sweep generator. Circuit simplicity and ordinary components provide a versatile test instrument.

how it works

The principle of the variable oscillator is based on a vacuum tube circuit originally described by K.C. Johnson in the April/May, 1949 issue of *Wireless World*. Although it has been revived in solid-state form since that time, the circuit never really caught on for reasons perhaps best expressed by Johnson himself: "It would appear," he said, "that most people do not believe that it could ever work."

The circuit is basically that of a Butler oscillator (**fig. 1A**) tuned to a frequency determined by L1 and C1. In the sweep oscillator, L1 is made electrically variable by what is known as "turns cancellation." This is achieved by coupling the coil to another coil, L2, which passes an RF current of opposite phase.

The magnitude of this out-of-phase current depends upon the imbalance of a differential amplifier (**fig. 1B**).

For example, when applying more negative voltage to its base, the gain of Q1 is reduced and Q2 is conse-

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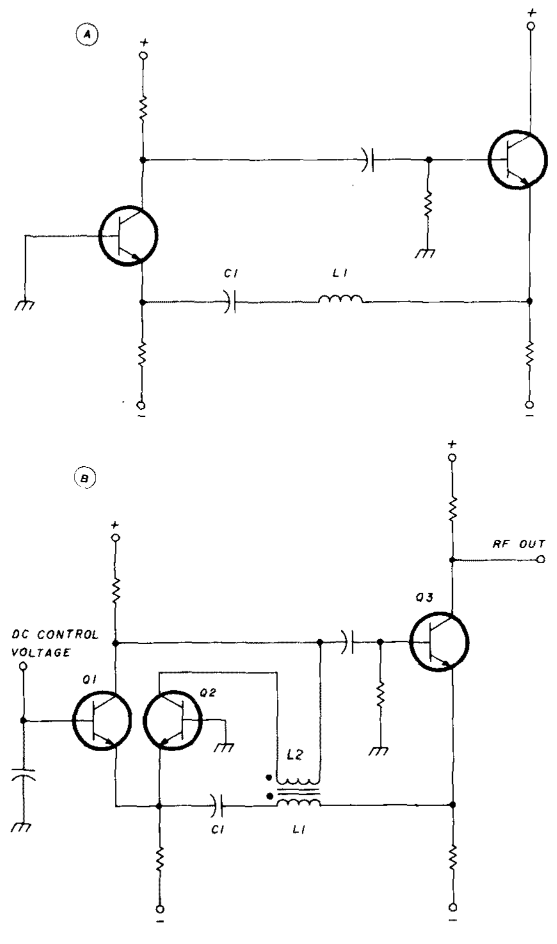


fig. 1. Heart of the sweep generator is based on the Butler oscillator, *A*. In the voltage-controlled oscillator, *B*, the effective coil inductance is controlled by "turns cancellation."

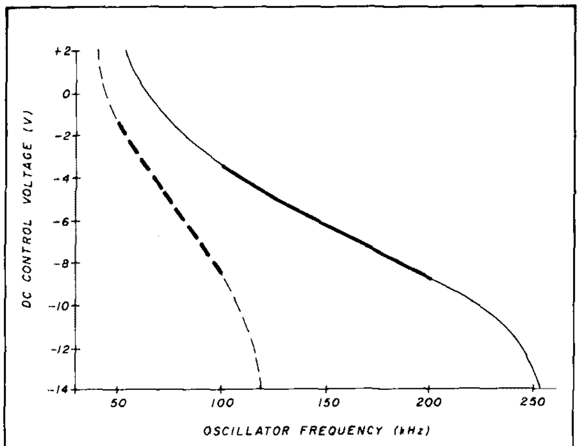


fig. 2. Oscillator basic frequency as function of DC control voltage, with C = 3750 pF. Dotted curve: C = 7500 pF.

quently increased, thus increasing the RF current through the coupling coil, L2.

As the direction of this RF current opposes the current through L1, the inductance of the frequency-determining L1 is effectively reduced, with the result that the oscillator frequency rises.

The more turns on L2, and the tighter the coupling between L2 and L1, the more frequency deviation can be accomplished. Therefore, the principle has nothing to do with the magnetic properties of the coil as such, and the obtainable frequency variation is greater with powdered iron or ferrite toroids than with ordinary air-wound coils, only because the core material permits the "cancellation" coil to assert a stronger influence on the tuning element.

After some experimenting it soon becomes clear that this is a remarkable circuit. Not only can frequency variations be made unusually large; the frequency linearity is also quite spectacular, as **fig. 2** shows. The total range of the oscillator covers frequencies up to a ratio as great as 1:5, including a range of 1:2 with excellent linearity.

Those who would like to experiment a little further with this "smart" circuit, may find it interesting to know that though the closely coupled coils are bound to introduce some stray capacitance, this does not prevent the circuit from working at much higher frequencies. Note that the RF voltages developed on the coils are of the same order: at the right-hand side they are connected to the same emitter follower. Indeed, the principle turns out to be useful for oscillators covering even the highest HF bands.

oscillator

The differential amplifier allows the tuning to be controlled from two independent sources. The base of Q1 is controlled by potentiometer R2 CENTER FREQ, while the base of Q2 receives its control voltage from the time-base generator (**fig. 3**). This provides a simple solution for making the sweep width symmetrical around the center frequency, regardless of amplitude. As the frequency/voltage sensitivity remains constant over the entire CENTER FREQ range, the output of the saw-tooth generator can be calibrated as Δ FREQ, which remains valid for all settings of CENTER FREQ.

A high-quality potentiometer is recommended for R2. In fact, it is a type with very thin and closely wound resistance wire; however, a good-quality, large-size carbon potentiometer would also do the job.

It is the oscillator's stability that determines the highest usable harmonic of the sweep generator. In practice this may be more than 10 MHz, providing that the following precaution is taken: the frequency stability can be improved considerably by joining Q1 and Q2 with "super-glue" and then wrapping the pair in

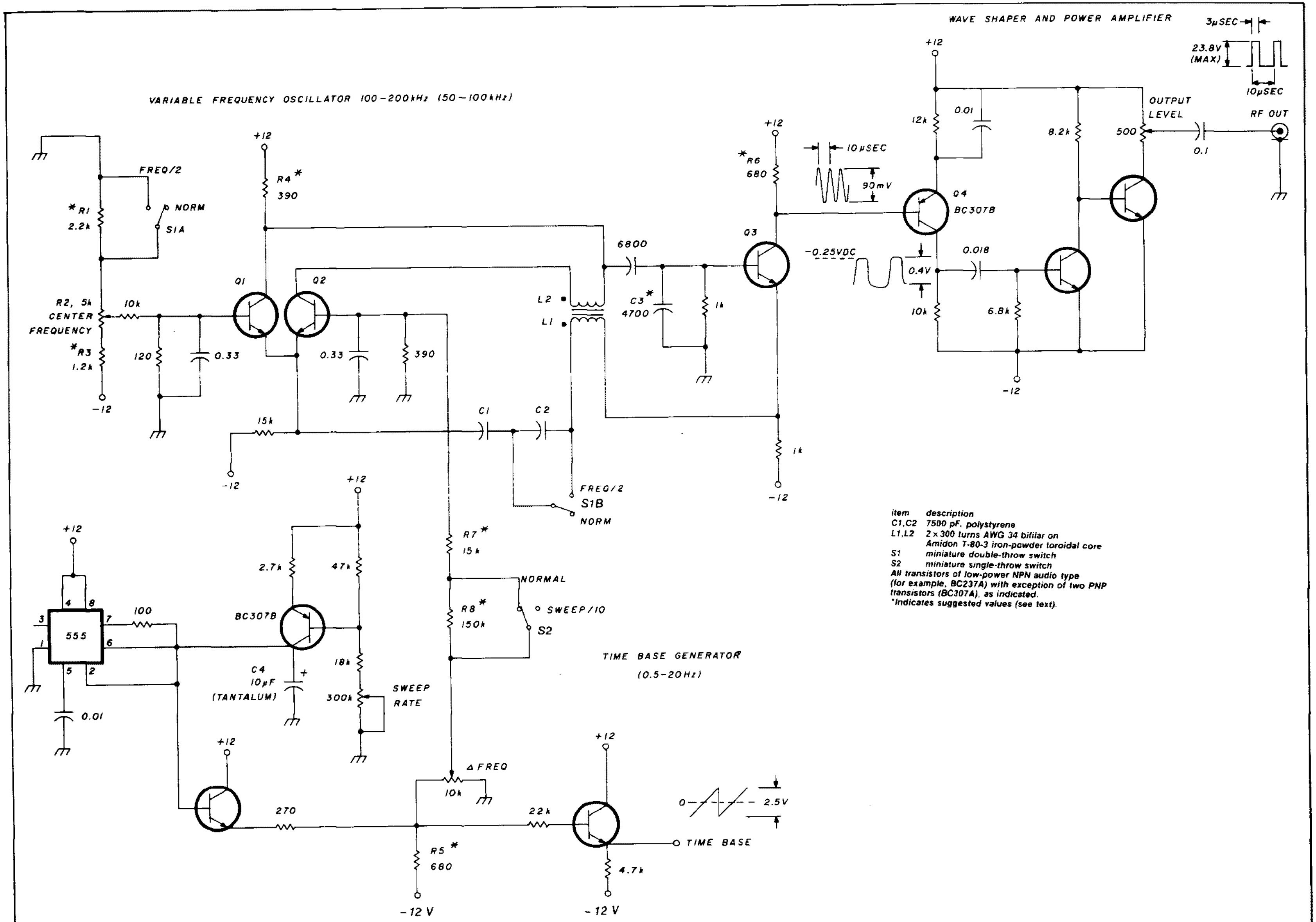


fig. 3. Schematic diagram of the compact IF sweep generator.

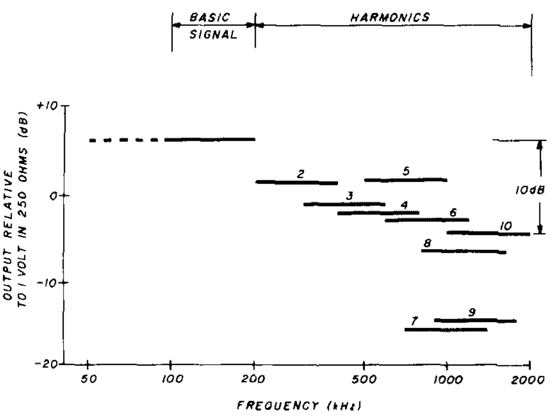


fig. 4. Effective output level of basic signal and its harmonics up to 2 MHz.

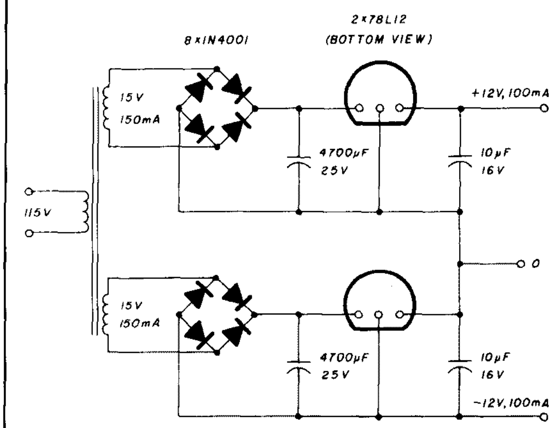


fig. 6. Power supply for sweep generator.

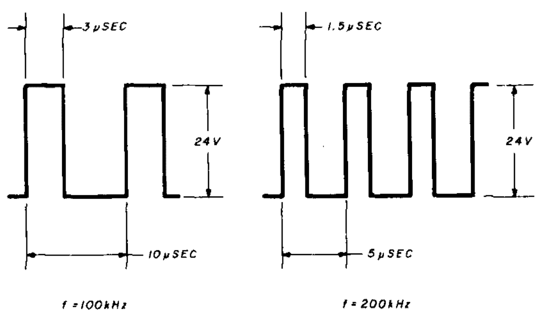


fig. 5. Output waveforms of the sweep generator. Left at 100 kHz, right at 200 kHz. As with varying frequency the duty cycle remains constant. The RF power level does not alter within the tuning range of each harmonic.

thin copper wire, and finally soldering the total into one solid metal blob. This assures that the two transistors will remain at virtually the same temperature, while their common, relatively heavy thermal mass prevents fast frequency drift.

In applications in which the sweep generator is used for sweeping filters at frequencies lower than 100 kHz, the basic range of the oscillator may be dropped to 50 to 100 kHz at the expense of a slight deterioration in frequency linearity. Double-throw switch S1 shorts one of the tuning capacitors and also shorts a resistor in series with the CENTER FREQ potentiometer. This enables the oscillator to work on half frequencies within the linear portion of the control range.

The calibration of both CENTER FREQ and Δ FREQ could be done very accurately by listening to the harmonics of the oscillator with a communication receiver.

time base

The sawtooth generator circuit enables sweep rates from twice a second, for slowly sweeping sharp-edged responses from crystal filters without the risk of ringing effects, to 20 Hz, permitting a flicker-free oscilloscope display.

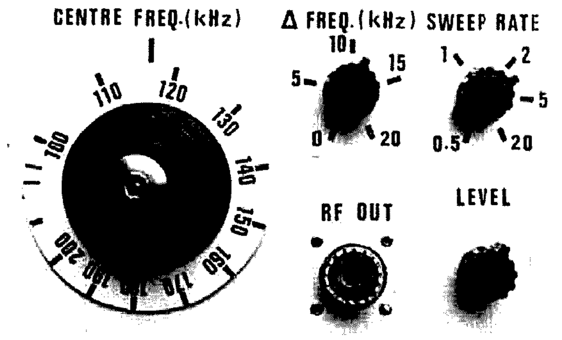
To keep the frequency deviations symmetrical around the center frequency, R5 may need some correction. Also the values given for R7 and R8 are approximate, as these resistors determine the Δ FREQ calibration.

When using a harmonic of the basic oscillator signal, not only does the frequency *increase*, but the indicated Δ FREQ value must also be equally multiplied. At higher frequencies the sweep width may consequently become too large and therefore impractical. Thus a provision was necessary for dividing the sweep width by a factor 10 in the form of R8. Normally this extra series resistor remains shorted by S2.

wave-shaping circuit

This circuit produces the harmonics of the oscillator signal. As a general rule, the widest harmonic spectrum may be created by generating the shortest pulses. However, the shorter the pulses, the less power each harmonic contains. After some experimenting (to compensate for my insufficient experience with Fourier-analysis techniques) I decided to aim for an output pulse with a duty cycle of about 30 percent. This is not too difficult to obtain and results in an overlapping spectrum of output signals, the levels of which remain — at least up to 2 MHz — within a 10-dB range (fig. 4). Only above 2 MHz does the effective output tend to drop beyond this range.

The signal strength of each individual harmonic is a different matter. It should remain constant during



IF sweep generator has a frequency range of 1:2 with excellent linearity.

the sweeping process. To make this possible, not only the amplitude but also the duty cycle of the rectangular waveform must remain constant, regardless of frequency.

A satisfactory sine wave can be obtained from the oscillator section by trimming R4 and C3. At the same time, a constant amplitude is maintained over the full frequency range. Q4 conducts mainly on the tops of this sine wave. These tops are then amplified and clipped, which results in a rectangular waveform (**fig. 5**). The surface of these pulses (amplitude X time) stays constant, regardless of frequency. This assures a constant output level of each harmonic.

The pulse characteristic of the output signal permits the use of an ordinary low-power transistor for the output stage. Functioning as an electrical switch, rather than an analog amplifier, the transistor must handle neither appreciable voltage nor current at any common moment in time. This explains how a perfectly cool little 200 mW transistor is capable of delivering the total output power (basic signal plus all harmonics together) of almost 1 watt.

The output impedance of the final stage depends on the setting of the OUTPUT LEVEL potentiometer. At maximum setting, the output impedance is equal to half the value of the potentiometer; that is, 250 ohms. Still, the sweep generator can be safely loaded with 50 ohms impedance without upsetting anything. Only an amplitude over 50 ohms will drop to one-fifth; that is, about 5 volts.

power supply

The power supply (both +12 volts and −12 volts at about 65 mA) may be very simple (**fig. 6**), using 78L12 voltage regulators. The hum level must be low, because any ripple on Q1 and Q2 control voltages causes frequency modulation. Although the toroidal coil L1/L2 is relatively insensitive to stray fields, it may nevertheless pick up hum from a nearby transformer, causing a rippling oscilloscope display. It is for this reason that the power supply has not been incorporated into the sweep generator itself, but is instead connected at the end of a 3-wire lead.

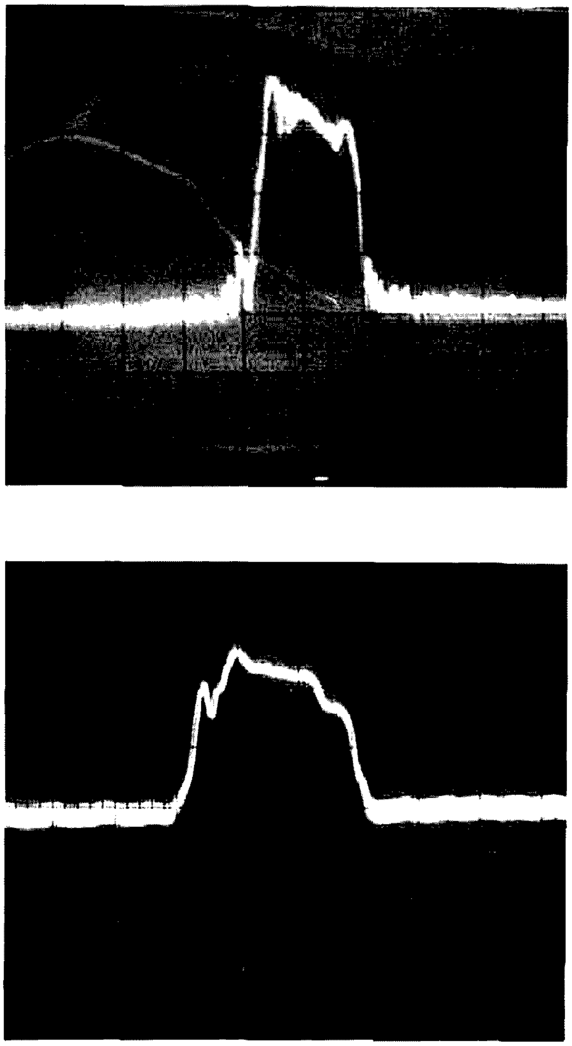


fig. 7. ***Top***: **Response of IF amplifier with 9 MHz crystal filter in HF transceiver. Sweep rate 4 Hz.** ***Bottom***: **poorly matched mechanical bandfilter in 455 kHz IF amplifier (Collins F455-J-31). Sweep rate 4 Hz.**

The RF detector for this sweep generator is a simple diode detector. Two examples using this detector are shown in **fig. 7**. However, the value of the measurement can be enhanced considerably by using a logarithmic detector instead. This permits an amplitude display on a decibel scale, over several decades, if necessary. (A simple version of such a logarithmic detector will be the subject of a forthcoming article in *ham radio*.)

ham radio

graphical selection
of mixer frequencies

See, at a glance, any spurious that might cause problems

Selecting the proper mixer frequencies can be a real problem. Often a lengthy trial-and-error procedure yields unsatisfactory results because of too many spurious signals in the passband. The graphical technique described here will deliver more accurate results in less time and with less difficulty. Some plotting is required, and a simple calculator will help with the math.

background

When two frequencies, f_1 and f_2, are combined in a mixer, the nonlinear action of the mixer produces a series of products that have the form:

$$P = Mf_1 + Nf_2 \tag{1}$$

where M and N are positive or negative integers, and P is the frequency of the combination. In ordinary mixer use, a bandpass or low-pass filter removes all but the desired product P, called the desired output frequency, or f_0. Generally, the larger M and N are, the smaller the amplitude of P. Then, too, the farther a particular P is from f_0, the less interference it will cause. One way to measure the frequency separation is to use the percentage separation, S, given by:

$$S = 100\,\frac{P - f_0}{f_0} \tag{2}$$

Now, if f_1 is always chosen as the smaller of f_1 and f_2, then the ratio f_1/f_2 can be given by:

$$f_1/f_2 = \frac{-S + 100(N-1)}{S - 100(M-1)} \tag{3}$$

where $f_0 = f_2 + f_1$

$$\text{or } f_1/f_2 = \frac{S + 100(1-N)}{S + 100(1+M)} \tag{4}$$

where $f_0 = f_2 - f_1$

Equations 3 and **4** are used to plot the spurious components, or spurs; **eq. 3** is used for sum spur charts, and **eq. 4** is used for difference spur charts. The case for N = M appears as a vertical line on the sum chart at S = 100(N − 1), and the case for N = −M appears as a vertical line on the difference chart, also at S = 100(N − 1).

conversion of fixed frequencies

Figure 1 shows a graph with several spur plots for f_1 = 3 MHz and f_2 = 18 MHz. Which ones need to be plotted? Well, that depends on how stringent your requirements are. Once you determine how low the spurs must be, you need plot only the spurs which

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table 1. Typical values of spurious levels in a double-balanced mixer. Values shown are in dB below desired output. Here, f_2 is the local oscillator frequency.

	f_2	$2f_2$	$3f_2$	$4f_2$	$5f_2$	$6f_2$
$6f_1$	90	90	90	90	90	90
$5f_1$	80	90	71	90	68	90
$4f_1$	90	86	80	88	85	86
$3f_1$	64	69	50	77	47	74
$2f_1$	73	74	70	71	64	69
f_1	0	35	13	40	24	45

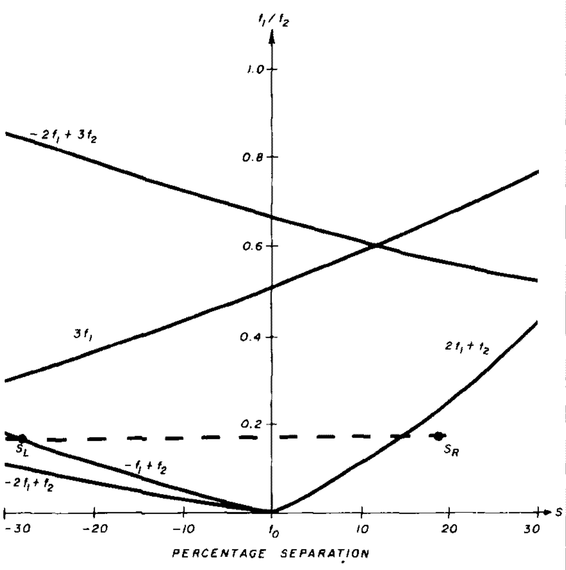


fig. 1. Spur chart for fixed-frequency conversion example: $f_0 = f_2 + f_1$.

might exceed that level. Some manufacturers provide tables of spur levels for different values of M and N.* **Table 1** shows some typical values.

To use a spur chart such as shown in **fig. 1**, you must somehow represent the bandwidth of the system. In **fig. 1**, this is done by plotting a dashed line representing an output frequency range for a given f_1 and f_2. The end points are denoted by S_L and S_R.

$$S_L = 100\frac{f_{OL} - f_O}{f_O} \tag{5}$$

and

$$S_R = 100\frac{f_{OR} - f_O}{f_O} \tag{6}$$

*Mini-circuits, P.O. Box 166, Brooklyn, New York 11235, and Watkins-Johnson Company, 3333 Hillview Avenue, Palo Alto, California 94306, for example, furnish tables of this kind.

where f_{OL} and f_{OR} are, respectively, the low and high ends of the output passband. S_L and S_R are the abscissa values; the ordinate value, f_1/f_2, is already known. Refer again to **fig. 1**. If, for example, an output passband of 10 MHz is chosen, from 15 MHz to 25 MHz, then $S_L = -28.6$ and $S_R = 19.0$. The ordinate is $f_1/f_2 = 3/18 = 0.17$. Note that two of the spurs intersect the bandwidth line, $-f_1 + f_2$ and $2f_1 + f_2$. This means that these two spurs are in-band harmonics. If the bandwidth was narrowed to, say, 5 MHz (18.5 MHz to 23.5 MHz), then the end points would be closer together and the spurs would be outside the passband. Note that if the slope of the filter is known, the attenuation of the out-of-band spurs can be calculated because **fig. 1** indicates how far out on the filter skirt the spurs are located. Of course, there is still the problem of the 18-MHz signal in the output. It, too, is a spur, and must be taken into consideration since it will be down probably no more than 50 dB.

conversion of bands of frequencies

Now that I've discussed a specific case with fixed frequencies, let's consider the general case — converting one band of frequencies to another band of frequencies where f_1, f_2, and f_0 have different bandwidths. Suppose you want to convert 300 MHz ±10 MHz to 200 MHz ±15 MHz by mixing with 100 MHz ±5 MHz. You have f_{1L} = 95 MHz, f_{1H} = 105 MHz, f_{2L} = 290 MHz, f_{2H} = 310 MHz, f_{0L} = 185 MHz, and f_{0H} = 215 MHz. First you'll outline the region you want to be free of spurs. (In general, this takes the shape of a hexagon with slightly curved sides. Since the curvature is slight, you can assume straight lines to ease the computations and still retain good accuracy.) This is done by using **eqs. 5** and **6** to calculate the corner points. Different combinations of the high and low extremes of f_1 and f_2 are used to find each particular f_0. Here's how to do it (remember — $f_0 = f_2 - f_1$ here).

Calculate: S_L using f_{1L} and f_{2H}
S_L using f_{1H} and f_{2H}
S_L using f_{1H} and f_{2L}
S_R using f_{1H} and f_{2L}
S_R using f_{1L} and f_{2L}
S_R using f_{1L} and f_{2H}

These calculations produce the numbers shown in **table 2**, and the six points 1 through **6** in **fig. 2**; the dashed lines define the desired spur-free area. For this combination of frequencies, two spurs cross the hexagon. Either of two things can be done to resolve this. You can change the frequencies or try a high-level mixer with reduced drive, which will result in fewer spurs.

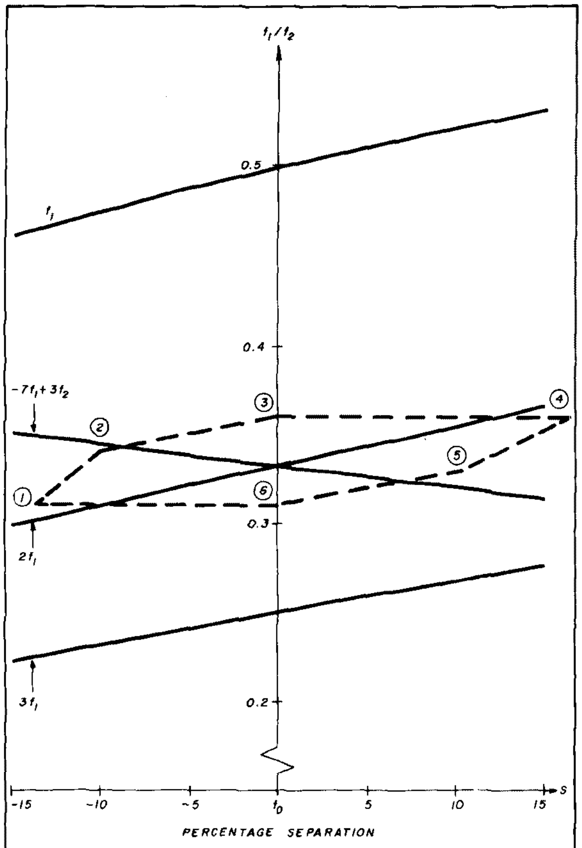


fig. 2. Spur chart for conversion of bands of frequencies — $f_0 = f_2 + f_1$. The circled numbers (1-6) are those referred to in the text and listed in *table 2*.

table 2. Values for points 1 through 6 in *fig. 2*.

point	S_L	S_R	f_1/f_2
1	−13.95	—	0.31
2	− 9.76	—	0.34
3	0.00	—	0.36
4	—	16.22	0.36
5	—	10.26	0.33
6	—	0.00	0.31

As a final example, let's try upconverting 28-32 MHz to 50-54 MHz by mixing with 22 MHz. Here, f_1 has zero bandwidth ($f_{1L} = f_{1H}$). This produces the dashed line (---) shown in **fig. 3**. Two things are evident: the hexagon is now a quadrilateral (because one of the frequencies, f_1, has zero bandwidth), and one of the spurs cuts through this quadrilateral. If you select a different combination — say, $f_1 = 22.5$ MHz and $f_2 = 27.5 - 31.5$ MHz — then you get a quadrilateral as shown by the dashed-and-dotted line (-·-) in **fig. 3**. This new area is spur-free so the filtered output will be free of spurs.

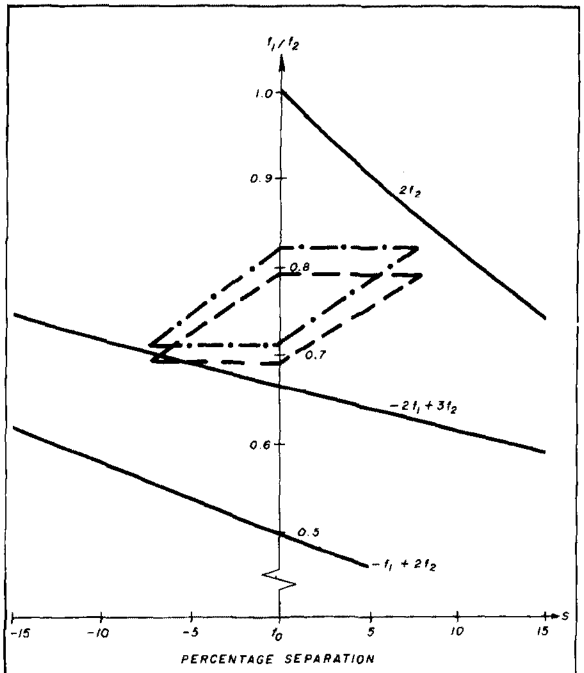


fig. 3. Spur chart with f_1 having zero bandwidth — $f_0 = f_2 + f_1$.

appendix

Equation 2 can be rewritten to give:

$$P = \frac{f_0 S}{100} + f_0 \tag{7}$$

For the case when $f_0 = f_2 + f_1$, inserting **eq. 1** in **eq. 7** yields $P = M(f_0 - f_2) + Nf_2$, or

$$f_2 = \frac{P - Mf_0}{N - M} \tag{8}$$

Equation 8 can be used in the expression for f_1/f_2 to eliminate f_2:

$$f_1/f_2 = \frac{f_0 - f_2}{f_0} = \frac{Nf_0 - P}{P - Mf_0} \tag{9}$$

Now, if **eq. 7** is inserted into **eq. 9**:

$$f_1/f_2 = \frac{Nf_0 - (f_0 S/100 + f_0)}{f_0 S/100 + f_0 - Mf_0}$$

$$= \frac{-S + 100(N - 1)}{S - 100(M - 1)} \tag{3}$$

If $f_0 = f_2 - f_1$, a similar procedure gives:

$$f_2 = \frac{P + Mf_0}{N + M} \tag{10}$$

$$f_1/f_2 = \frac{P - Nf_0}{P + Mf_0} \tag{11}$$

and

$$f_1/f_2 = \frac{S + 100(1 - N)}{S + 100(1 + M)} \tag{4}$$

reference

1. M.Y. Huang et. al., "Select Mixer Frequencies Painlessly," *Electronics Design*, No. 8, April 12, 1976, pages 104-109.

ham radio

Morse code computer tutor

Let your VIC-20 help you upgrade

For many Amateurs, it isn't theory or rules and regs that make upgrading difficult — *it's the code*. Jumping from a plateau of 10 words per minute or 18 words per minute seems impossible.

Satisfying these code requirements is even more difficult if you don't get enough practice. If CW isn't your favorite mode of operation, you're not likely to push yourself to practice on the air. Practice with code tapes can be productive — until you start memorizing them. And W1AW's code practice, while useful, may not always match your schedule or be strong enough to copy.

With this program (**fig. 1**), you can turn your VIC-20 personal computer into your own personal Morse code computer tutor. This tutor will give you all the practice you ask for at any time, at any speed, and always at Q5 conditions.

what the program does

The computer tutor sends groups of 20 random Morse code characters at any user-selectable speed starting at 4 words per minute. You type in the characters as they are sent. After the twentieth character the tutor identifies the random characters it has sent, and then lists your responses and your score. A particular advantage of this system of practice is that it forces you to copy Morse code with a keyboard; without a keyboard, few people can transcribe code at speeds greater than 25 WPM.

how the program works

After printing a sign-on message the program inputs the user-selected speed in **line 5**. **Lines 6** through **9** initialize certain variables such as character counter, correct response counter, and character strings. **Lines 20** through **91** are subroutines that store all the dot/dash patterns and their printable equivalents.

The selection of the random character and its conversion into Morse code begins at **line 100**. Here the program generates a random number between 1 and 36 inclusive. **Lines 110** through **120** use the random number to select one of 36 characters — the letters

```
1 REM MORSE CODE COMPUTER-TUTOR
  BY LAWRENCE G SOUDER, N3SE
2 PRINT "▮I WILL SEND A GROUP OF TWENTY CHARACTERS.
  FOR  SCORE HIT RETURN."
3 PRINT "HOW MANY WORDS PER    MINUTE DO YOU WANT?"
5 INPUT V
6 S = 27-V
7 C=0
8 Z=0
9 K$="": T$=""
10 GOTO 100
20 C$=".-A"
21 RETURN
22 C$="-...B"
23 RETURN
24 C$="-.-.C"
25 RETURN
26 C$="-..D"
27 RETURN
28 C$=".E"
29 RETURN
30 C$="..-.F"
31 RETURN
32 C$="--.G"
33 RETURN
34 C$="....H"
35 RETURN
36 C$="..1"
37 RETURN
38 C$=".---J"
39 RETURN
40 C$="-.-K"
41 RETURN
42 C$=".-..L"
43 RETURN
44 C$="--M"
45 RETURN
46 C$="-.N"
47 RETURN
48 C$="---O"
49 RETURN
50 C$=".--.P"
51 RETURN
52 C$="--.-Q"
53 RETURN
54 C$=".-.R"
55 RETURN
56 C$="...S"
57 RETURN
58 C$="-T"
59 RETURN
60 C$="..-U"
61 RETURN
62 C$="...-V"
63 RETURN
64 C$=".--W"
65 RETURN
66 C$="-..-X"
67 RETURN
68 C$="-.--Y"
69 RETURN
70 C$="--..Z"
71 RETURN
72 C$=".----1"
73 RETURN
74 C$="..---2"
75 RETURN
76 C$="...--3"
```

fig. 1. N3SE program for Morse code training and practice on the VIC-20.

By **Lawrence G. Souder, N3SE**, 4539 Manayunk Avenue, Philadelphia, Pennsylvania 19128

```
77 RETURN
78 C$="....-4"
79 RETURN
80 C$=".....5"
81 RETURN
82 C$="-....6"
83 RETURN
84 C$="--...7"
85 RETURN
86 C$="---..8"
87 RETURN
88 C$="----.9"
89 RETURN
90 C$="-----0"
91 RETURN
100 R=INT(RND(1)*36)+1
103 IF C=20 GOTO 500
105 IF R > 21 GOTO 120
110 ON R GOSUB 20,22,24,26,28,30,32,34,36,38,40,42,
    44,46,48,50,52,54,56,58,60
115 GOTO 130
120 ON R-21 GOSUB 62,64,66,68,70,72,74,76,78,80,82,
    84,86,88,90
130 T$=T$+RIGHT$(C$,1)
131 C=C+1
134 C1$=RIGHT$(C$,1)
135 GOTO 200
140 GOTO 100
200 L=LEN(C$)
205 L=L-1
210 N=I
220 M$=MID$(C$,N,1)
230 N=N+1
240 GOSUB 400
250 IF L>=N GOTO 220
260 FOR X=1TO10*S:GET L$
262 IF L$<>""THEN K$=K$+L$
265 NEXT X
270 GOTO 100
400 S1=S
410 IF M$="-" THEN S1=7*S
420 POKE 36878,15
430 POKE 36875,240
440 FOR X=1TOS1*2: NEXTX
450 POKE 36878,0
460 FOR X=1TOS:GET L$
464 IF L$<>""THEN K$=K$+L$
468 NEXT X
470 RETURN
500 INPUT R
505 FOR X=1 TO 20
510 IF MID$(T$,X,1)=MID$(K$,X,1) THEN Z=Z+1
520 NEXT X
525 PRINT "♥"
527 PRINT
530 PRINT T$
540 PRINT K$
545 PRINT
550 PRINT "YOUR SCORE IS ";100*(Z/20);"%"
560 PRINT "AGAIN AT SAME SPEED?"
570 INPUT R$
580 IF R$="Y" GOTO 7
582 PRINT "AT DIFFERENT SPEED?"
584 INPUT R$
586 IF R$="Y" GOTO 3
590 END
```

A through Z and the numerals 0 through 9. For example, from the random number 7 these lines will select the character "G," which is held in **line 32**.

Before the dot/dash patterns are sent in Morse code, **line 130** separates the printable character from its Morse elements and saves it to be printed later. For example, it separates the letter "G" from the string "— —.G". **Lines 131** through **265** take each dot and dash and call up a subroutine to output a tone of the proper duration. This subroutine is in **lines 400** through **470**.

*The heart-shaped graphic character in lines 2 and 525 is used to clear the screen in the VIC-20.

Two FOR-NEXT loops produce the delays for the dots and dashes and for the spaces between dots, dashes, and characters. During these delay loops the program inputs any response from the user. After 20 characters have been sent, **lines 500** through **550** compare the random characters sent with the responses typed in and print out both strings of characters along with the user's score. Then **lines 560** through **586** ask the user whether more practice is desired at the same speed, at a different speed, or not at all.

modifying and adapting the program

You might improve the program by adding some of the other characters such as punctuation. Do this by adding character string statements after **line 91**. Then extend the range of the random number generator by the same amount. For example, if you want to add a comma, add these lines after 91: 93 C$ = "— —..— —,"; 94 Return. Then change the random number generator in **line 100** to: 100 R = INT(RND(1)*37) + 1 to make its range 1 through 37 inclusive.

Now there's no excuse to put off upgrading. With this computer tutor you can get as much code practice as you need at any time — and improve your typing skills, too.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

As this issue went to press, U.S. Amateurs received FCC approval to operate on the new 12-meter band (24.89-24.99 MHz) as of June 22. Unfortunately, we're approaching a sunspot minimum and the band will be relatively worthless for long-distance skip communication. But all is not lost — good contacts can be had by sporadic-E skip, and once in a while an unusual burst of activity from the sun will open the band for DX for a few hours. In any event, it's a good idea to get on the band and enjoy this new chunk of spectrum as soon as operation is authorized.

I've monitored the band for years and have heard plenty of DX when conditions were good. Over 40 countries are licensed for operation on the 12-meter band. See how many of them you can hear and work! Even at the low point of the sunspot cycle, the north-south path isn't bad, and you should be able to work some South American Amateurs when the band opens up.

antennas for 12 meters

Amateurs who have an "all-band" antenna with a tuner at the operating position can get on the band immediately. Others will have to improvise. One quick way to get on the air is to string up a dipole or inverted-V to your tower and feed it with a separate coax line as shown in **fig. 1**. The higher you can get it in the air, the better the results.

Another easy-to-erect antenna is a ground plane **(fig. 2)**. As in the case of the dipole, the higher you can erect it, the better it will work. If you have time to build a beam for the band, **fig. 3** provides dimensions for a quad antenna and **fig. 4** provides Yagi information.

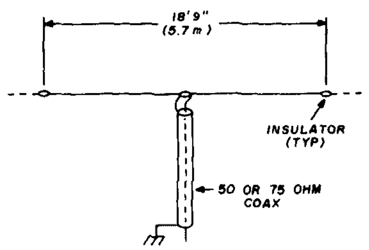


fig. 1. Get on the new 12-meter band quickly with this dipole or inverted-V antenna.

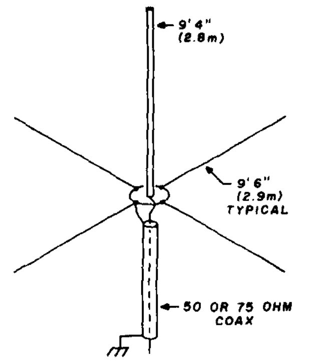


fig. 2. Simple ground plane for the new 12-meter band. Vertical portion is made of 3/4-inch aluminum tubing. Radials are No. 14 wire.

the G8PO "JAWS" antenna

G8PO has discussed an interesting variation of the quad loop that has provided superior results on 7 MHz **(fig. 5)**.[1] Mast height is less than that required for the conventional loop as the bottom portion is bent out of the vertical plane. The antenna is fed at one corner by a gamma match system to provide a good match to a coax line. Antenna polarization is vertical.

The lower portion of the loop has three conductors in parallel, running nearly horizontal to the ground. Tests indicated improvement in performance over the England-New Zealand test path and some front-to-back signal discrimination became apparent.

Checks against a conventional loop over the same path showed that the regular loop was consistently weaker during many contacts. The forward gain of the JAWS antenna was estimated to be 3 dB or better, and the front-to-back ratio was about 6 dB.

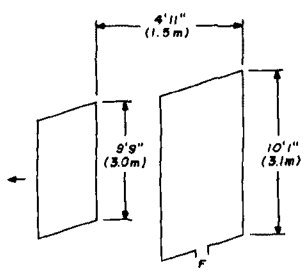


fig. 3. Two-element quad for the 12-meter band. Driven element is fed at point *F* with 50 or 75-ohm coax line.

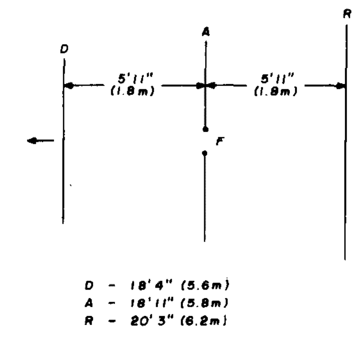


fig. 4. Dimensions for 12-meter beam made out of 1-inch tubing with 7/8-inch tip sections. Feedpoint resistance at *F* is about 20 ohms.

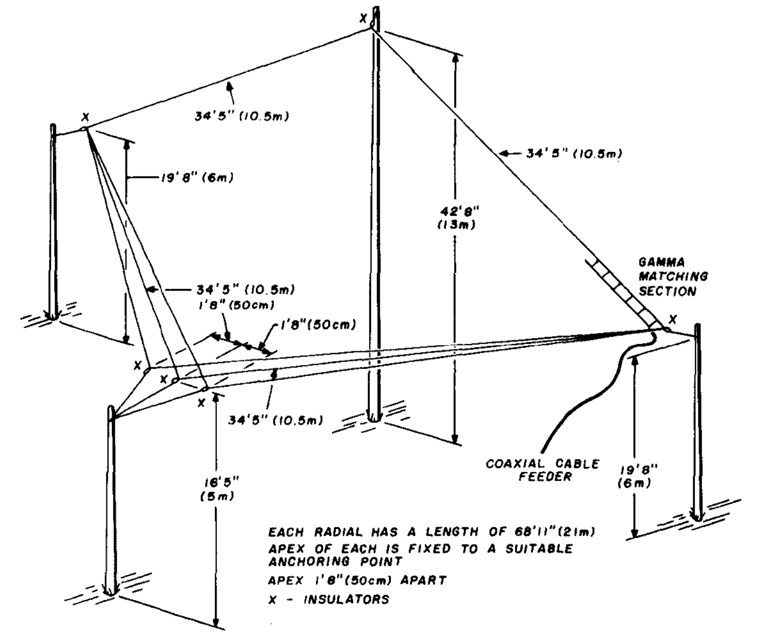


fig. 5. The G8PO "JAWS" Antenna for 40 meters. Directivity is to the left through the apex of the three parallel-connected wires. (*Figure courtesy RSGB.*)

The gamma match is made of wire, with the gamma section measuring about 6.5 feet (1.98 meters) long and spaced away from the loop wire about 4 inches (10 cm). The gamma capacitor is 200 pF.

the multiband antenna

Independent experimenters have discovered that altering the shape of a driven element can change the harmonic resonance without appreciably altering the fundamental resonant frequency of the antenna. This is a good technique to use for a two-band antenna. A typical linear element of uniform diameter, unfortunately, does not exhibit resonance on the exact harmonic frequencies because of end effects. A 7-MHz dipole, for example, is *not* resonant in the 21-MHz band.

However, by changing the shape of the element, the third harmonic resonant frequency can be lowered without changing the fundamental frequency to any great extent. The principle is illustrated in **fig. 6**. The vertical antenna *element A* exhibits a quarter-wave resonance at 3.6 MHz. By formula, the antenna is 65 feet (19.81 meters) high. The third harmonic resonance, by formula, falls at 11.6 MHz. The actual third harmonic of 3.6 MHz, however, is 10.8 MHz. Thus there is a difference of 800 kHz between the actual third harmonic of the fundamental frequency and the third harmonic resonance of the vertical.

If the vertical resonance at the third harmonic region could be "pulled" down to 10.1 MHz, then the antenna could operate in the 30-meter ham band (10.1 to 10.15 MHz). Can this be done without disturbing the resonance in the 80-meter band?

Figure 6B shows the technique to accomplish this. The antenna element is made "fatter" near the area of maximum third harmonic voltage. This provides additional capacitance to ground at this frequency. On the fundamental frequency, the voltage is much lower at this point in the antenna element and the capacitive effect to ground is much less. In this manner the third harmonic resonance frequency is lowered without too much effect on the fundamental frequency.

Shown in **fig. 7** is an antenna developed by K6KBE for two-band operation. A three-legged tower having a very thin upper portion and a tapered lower section, this antenna shows resonance on both the 80 and 160-meter bands **(fig. 8)**.

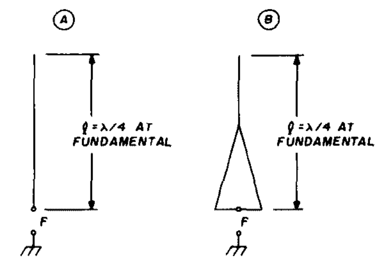


fig. 6. Antenna *A* shows classic response on fundamental and third harmonic. Antenna *B* exhibits lower-than-normal resonance on third harmonic while fundamental resonant frequency remains virtually unchanged.

The tower is 132 feet (40.2 meters) high with a base 12 feet (3.66 meters) on a side. The design frequencies are 1.85 and 3.7 MHz. The SWR across the 160-meter band is less than 1.8:1 at the band edges and below 1.5:1 from 3.5 to 3.9 MHz, rising to 2.3:1 at 4 MHz.

The top 44 feet (13.41 meters) of the antenna consists of a flexible aluminum whip, 2 inches (5.08 cm) in diam-

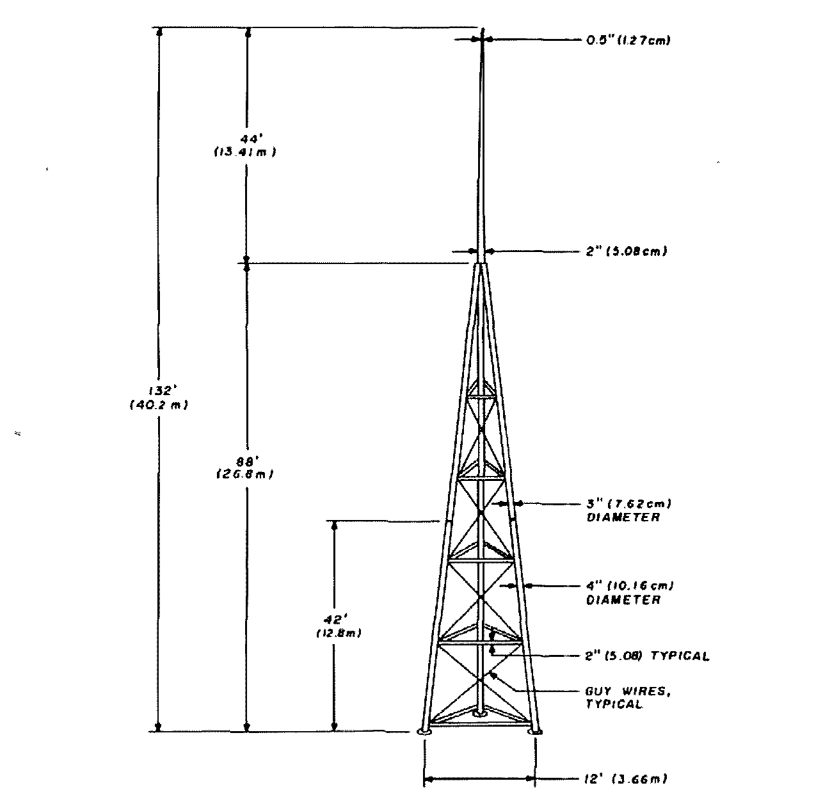


fig. 7. Details of 160/80-meter vertical antenna at K6KBE. Three-legged tower is 88 feet (36.88 meters) high with 44 foot (13.41 meter) whip on top.

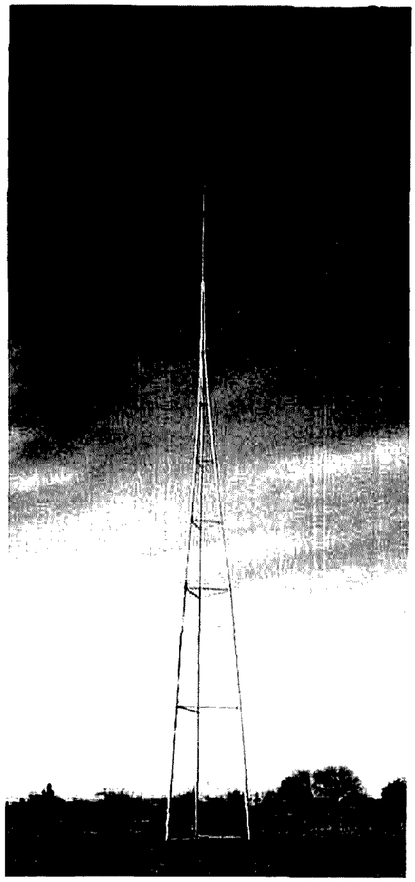


fig. 8. The self-supporting 160/80-meter vertical antenna at K6KBE.

eter at the butt, tapering to 0.5 inch (1.27 cm) at the tip. The whip is actually 2 feet (61 cm) longer than this, with the extra length forming the joint to the main tower, which is 88 feet (26.82 meters) high.

The bottom 42 feet (12.8 meters) on the tower is made of aluminum tubing 4 inches (10.16 cm) in diameter, with a 0.093 inch wall thickness. The top portion [to the 88 foot (26.8 meters) level] is made of 3-inch (7.62 cm) diameter tubing having a 0.063 inch wall. The cross-guys are made of 0.25 inch (1.27 cm) aircraft cable. Turnbuckles permit the assembly to be tightened by the assembler until a very rigid structure is achieved.

Anyone who has heard K6KBE's signal on 80 or 160 meters knows this antenna works!

speaking of radials . . .

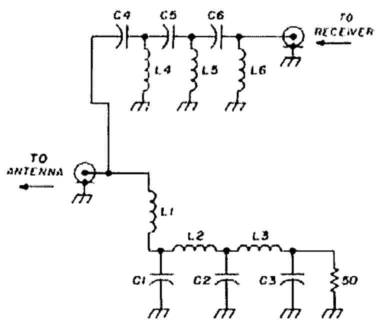
I just got a note from WA6BAN telling me more about his experiments with his 40-meter vertical ground plane antenna. He put it up with three radials, setting the base of the antenna a few feet above ground level. After he achieved resonance, he measured the feedpoint resistance with a General Radio RF Bridge. The result was about 58.3 ohms. He added a few more radials and the feedpoint resistance dropped to 53 ohms. Three more radials brought the resistance down to 51 ohms. Finally, he added more radials until he had eleven, and the feedpoint resistance dropped to 45 ohms.

His conclusion was that when the ground plane antenna is mounted close to the surface of the ground, you need "a lot more" than eleven radials to approximate a feedpoint impedance of 36 ohms. W2FMI, in his classic *QST* series on ground plane antennas,[2] came to the conclusion that sixty radials were required when they were laid on the surface of the ground.[3] When elevated radials are installed a few feet above the ground, it is possible that fewer will do the job. The correct number seems to be between eleven and sixty! (Anybody out there have a closer "fix" on this?)

broadcast filter for 160 or 80 meters

If you live in a residential or urban area, you can experience severe crosstalk and overload problems from local broadcast stations if you attempt to operate on 160 or 80 meters. (A friend of mine, located a few miles away from a local broadcast station, measured over 4 volts of RF pickup on his 80-meter vertical antenna. It completely locked up his receiver.)

Designed by K6KBE, the filter shown in **fig. 9** is an adaptation of the



Except as indicated, decimal values of capacitance are in microfarads (μF); others are in picofarads (pF); resistances are in ohms. k = 1,000 M = 1,000,000

	160M	80M
C1	3385	1862
C2	2318	1275
C3	500	275
C4	1242	683
C5	1242	683
C6	2554	1404
L1	7.5	4.1
L2	7.5	4.1
L3	3.64	2.0
L4	2.75	1.5
L5	4.01	2.2
L6	18.63	10.2

C - pF L - μH

fig. 9. Broadcast filter for 160 or 80 meters.

absorption filter originally used where suppression of harmonic energy is desired. In its original configuration, there are two complementary filters consisting of a high pass section terminated in a resistor and a low pass section to pass the desired signal. In this case, the reverse idea is used so that all energy *below* cutoff is routed to a dummy load while all energy above is allowed to pass.

The cutoff frequency for the 160-meter filter is 1.65 MHz; for the 80-meter filter, it's 3 MHz.

references

1. This material is extracted from "The G8PO JAWS Antenna," by Cdr. J.E. Ironmonger, G8PO, *Radio Communication*, November, 1984, pages 954-957. (Don't ask me what JAWS stands for — I can't figure it out either!)
2. Jerry Sevick, W2FMI, "The ground-image Vertical Antenna," *QST*, July, 1971, page 16.
3. For more information on ground radials and verticals in general see the K2BT series of articles on phased arrays, *ham radio*, May, June, July, October, December, 1983 and May, 1984 — Ed.

ham radio

designing Yagis
with the Commodore 64

See how gain and F/B can remain high across the entire band

In 1980, *ham radio* published a series of articles by W2PV on Yagi design.[1] As an avid DXer and a member of Potomac Valley Radio Club, I'd crossed paths with W2PV under contest conditions many times. He always had a fantastic signal, and after reading his series I realized why. W2PV was using superior antennas, all positioned at the correct height above ground.

Through computer research, W2PV showed that merely stating the mechanical length of an element of a Yagi beam does not actually reflect its true resonant frequency, especially when tapered elements are involved. He has also showed that tuning the reflector 5 percent below and the director 5 percent above the band center makes a poor beam. Commercial antenna specifications usually indicate maximum gain, maximum front-to-back ratio, and an SWR curve. But what they don't tell you is that the maximum gain and maximum front-to-back ratio do not occur at the same frequency, nor do they occur in the band center. They also don't tell you that as you tune away from these frequencies, the gain and maximum front-to-back ratio can fall off rapidly, indicating that although you may have 8 dB of gain at some frequency, you may only have 3 dB of gain near the band edges, depending on the tuning of the parasitic elements.

W2PV demonstrated that gain is primarily dependent on boom length and not on the number of elements on the boom, especially in the 1/4 to 3/4 boom length range. In addition, a naturally high front-to-back ratio occurs on a 1/4 wavelength boom and odd multiples thereof. He carefully designed his Yagis so that they would maintain a high gain and high front-to-back ratio over a 4 percent bandwidth, which will cover all of 15 and 20, most of 40, and a good portion of 10 meters. In order to do this the parastic elements must be tuned closer to the driven element reducing the SWR bandwidth. This is a small price to pay for maintaining almost maximum gain over the whole band. For example, the reflector of his three-element monobander is tuned 1.7 percent below the central design frequency and the director is tuned 4.2 percent above. He also proved that the resonant frequency of a tapered element could be very accurately determined.

About a year ago I wrote a program in BASIC using the formulas and data in the W2PV articles. With it you can turn your commercial antennas or scrap aluminum tubing into W2PV super beams. The program (**fig. 1**) is in simple BASIC with no peek or poke statements, and runs on the Commodore 64 computer.

program description

Lines 100-450 are a brief history of the program and instructions to the user. The type of beam you are designing is entered in **lines 460-500**. The actual lengths and diameters of each element are entered on **lines 550-620**, while the subroutine **lines 5000-5040** prints the inputs to the screen. **Lines 2000-2040** place data into the *A* array and if the average diameter calculated on **line 670** is 0.875 the date in *A* array is dumped into *R* array. If the average diameter is not 0.875, then new data is calculated in **lines 2100-2200** and placed into *B* array. Then *B* array is dumped into *R* array via **lines 2210-2220**. The element half-length that you are trying for per the central design frequency you entered earlier

By **Alan Hoffmaster, WA3EKL,** 929 Andrews Road, Glen Burnie, Maryland 21061

fig. 1. WA3EKL program for designing 10, 15, and 20-meter Yagis with the Commodore 64.

```
100 PRINT"MONO BAND ANTENNA DESIGN"
102 DIMC$(3),B$(4)
104 PRINT:PRINT"DO YOU WISH INSTRUCTIONS(Y/N)";:INPUTC$
106 IFC$="N"THEN450
108 PRINT
110 PRINT"THIS PROGRAM WAS WRITTEN USING THE FORMULAS PUBLISHED IN 1980 HAM RADIO"
112 PRINT"MAGAZINE AND WRITTEN BY W2PV WHO PROVED MATHEMATICALLY AND EXPERMENTAL
114 PRINT"THAT ANTENNA ELEMENT RESONATE FREQUENCY LENGTH IS NOT A SIMPLE"
116 PRINT"CALCULATION ESPECIALLY WHEN TAPERED ELEMENTS ARE INVOLVED."
120 PRINT
130 PRINT"MECHANICAL RECOMMENDATIONS"
140 PRINT"20 METER BOOM SHOULD BE 3 OR 4 INCH DIAMETER ALUMINUM IRRIGATION"
150 PRINT"PIPE DEPENDING ON LENGTH."
160 PRINT"ELEMENTS SHOULD BE TAPERED SEGMENTS FROM 1 1/4 INCH TO 1/2 INCH"
165 PRINT"DIAMETERS OF 6061 T6 SEAMLESS TUBING OR EQUIVALENT."
182 DIMI$(3)
185 PRINT:PRINT"TO CONTINUE INSTRUCTION PRESS RETURN";:INPUTI$
190 PRINT:PRINT"10-15 METER BOOMS SHOULD BE MADE OF 2 INCH ALUMINUM IRRIGATION PIPE."
200 PRINT"ELEMENTS SHOULD BE TAPERED FROM 7/8 INCHES TO 5/8 INCH DIAMETERS OF"
205 PRINT"6061 T6 SEAMLESS TUBING OR EQUIVALENT."
210 PRINT:PRINT"ELECTRICAL RECOMMENDATIONS."
215 PRINT"FOR MAXIMUM FRONT TO BACK RATIO USE A 1/4 WAVELENGTH BOOM ON A 3 ELEMENT"
220 PRINT"BEAM AND 3/4 WAVELENGTH BOOM ON A 4,5 OR 6 ELEMENT BEAM."
225 PRINT"EQUALLY SPACE ALL ELEMENTS ALONG THE BOOM REQUARDLESS OF LENGTH."
230 PRINT"FOR A GOOD 2 ELEMENT BEAM USE A BOOM LENGTH OF .150 WAVELENGTH."
240 PRINT"FOR A GOOD 3 ELEMENT BEAM USE A BOOM LENGTH OF .300 WAVELENGTH."
245 PRINT:PRINT"TO CONTINUE INSTRUCTIONS PRESS RETURN";:INPUTI$
250 PRINT:PRINT"IMPORTANT ALL TOTAL LENGTH DIMENSIONS WILL BE ELEMENT HALF LENGTH"
255 PRINT"MEANING THE LENGTH OF THE ELEMENT MEASURED FROM THE BOOM OUT TO THE"
260 PRINT"TIP OF THE ELEMENT. ENTER ELEMENT SEGMENT LENGTHS AND DIAMETERS JUST"
265 PRINT"AS THEY WILL BE ASSEMBLED STARTING WITH THE LARGEST DIAMETER SEGMENTS"
270 PRINT"(WHICH CROSSES OVER THE BOOM) TO THE SMALLEST DIAMETER SEGMENT AT THE"
275 PRINT"ELEMENT TIP. THE OBJECT IS TO ADJUST THE ORIGINAL SEGMGNT LENGTHS SO"
280 PRINT"THAT THE TOTAL EQUIVALENT LENGTH IS EQUAL TO OR VERY CLOSE TO THE"
285 PRINT"NORMALIZED LENGTH FOR THE PARTICULAR ELEMENT YOUR ARE WORKING ON."
290 PRINT"TO CONTINUE PRESS RETURN";:INPUTI$
295 PRINT"WHEN THE COMPUTER ASKS YOU FOR THE INFORMATION ENTER IN THE INFORMATION"
300 PRINT"THEN PRESS RETURN. NOTHING WILL HAPPEN UNTIL YOU PRESS RETURN. IF"
305 PRINT"NOTHING HAPPENS IMMEDIATELY, WAIT. THE COMPUTER MAY BE THINKING."
310 PRINT:PRINT"TO CONTINUE PRESS RETURN";:INPUTI$
450 PRINT:PRINT"ENTER CENTRAL DESIGN FREQUENCY IN MHZ";:INPUTCDF
460 RL=11811.024/CDF:REM RESONATE LENGTH IN INCHES
470 PRINT"ENTER NUMBER OF ANTENNA ELEMENTS 2,3,4,5 OR 6";:INPUTP
480 IFP<2ORP>6THEN470
490 PRINT"ENTER BOOM DIAMETER";:INPUTBD
496 BC=(BD/2)*0.0625:REM BOOM CORRECTION
500 PRINT"ENTER NUMBER OF ANTENNA SEGMENTS";:INPUTN
505 X=0
510 DIML((N-1)),D((N-1)),F((N-1))
520 FORE=0TO(N-1)
530 L(E)=0:D(E)=0:F(E)=0
540 NEXTE:REM CLEAR ARRAYS
550 PRINT:PRINT"ENTER(IN INCHES) LENGTH OF SEGMENT";:INPUTA
560 L(X)=A
570 PRINT"ENTER(IN INCHES) DIAMETER OF SEGMENT";:INPUTB
580 D(X)=B
590 GOSUB5000
600 X=X+1:REM STEPPING DATA INTO L ARRAY
610 IFX=NTHEN630
620 GOTO550:REM ENTERING IN SUCCESSIVE LENGTHS AND DIAMETERS
630 GOSUB4000
640 FOREE=0TO(N-1)
650 TD=TD+D(EE)
660 NEXTEE
670 AD=TD/N
680 GOSUB2000
690 GOSUB3000
700 T=0:G=0:REM CLEAR TOTAL
710 FORY=0TO(N-1)
720 T=T+L(Y):REM ADD SEGMENT LENGTHS
730 PRINTL(Y);"INCHES LONG",D(Y);"INCHES DIA."
740 NEXTY
750 PRINT:PRINTT"INCHES=TOTAL ORIGINAL LENGTH"
770 O1=0:O2=0:W=0
780 R1=AD/(2*RL):REM = RADIUS OF CENTER SEGMENT
790 D1=D(G)
800 R2=D1/(2*RL):REM = RADIUS OF SUCCESSIVE ELEMENTS
810 G=G+1:REM STEPPING THRU SEG. DIAMETERS.
820 K1=1/R1
830 K2=1/R2
840 M=((43.03*(LOG(K2)/LOG(10))-32))/((43.03*(LOG(K1)/LOG(10))-32))
850 O2=(L(W)*1.57079632)/T+O1:REM O2 IN RADIANS
860 FO=(SIN(2*O2)-SIN(2*O1))/((2*O2)-(2*O1))
870 O1=O2
880 SA=((M+1/M)/2+(M-1/M)*FO/2)*L(W)
890 F(W)=SA:REM PLACE EQUIVALENT LENGTHS INTO F ARRAY
900 W=W+1:REM STEPPING THRU F ARRAY
910 IFW=NTHEN930
920 GOTO780:REM LOOP TO CALCULATE EACH SEGMENT LENGTH
930 S=0:REM CLEAR EQUIVALENT LENGTH TOTAL
940 FORZ=0TO(N-1)
950 S=S+F(Z):REM ADD EQUIVALENT LENGTHS IN F ARRAY
960 NEXTZ
970 PRINT:PRINTS"INCHES=TOTAL EQUIVALENT LENGTH
972 PRINT:PRINT"HARDCOPY(Y/N)"
973 GETZ$:IFZ$=""THEN973
974 IFZ$="Y"THENGOSUB5090
980 PRINT"DO YOU WISH TO CHANGE A SEGMENT LENGTH(Y/N)";:INPUTB$
990 IFB$="N"THEN1170
1000 PRINT"ENTER NUMBER OF SEGMENTS YOU WISH TO CHANGE COUNTING FROM THE BOOM OUT"
1010 INPUTV
1100 IFV<1ORV>NTHEN1000
1130 PRINT"ENTER NEW SEGMENT LENGTH(IN INCHES)";:INPUTU
1140 L(V-1)=U
1150 GOSUB3000
1160 GOTO700
1170 PRINT"{CLR}":REM CLEAR SCREEN
1180 PRINTP;"ELEMENT BEAM AT ";CDF;"MHZ"
1190 PRINT
1200 FORTT=0TO(N-1)
1210 PRINTL(TT);"INCHES LONG",D(TT);"INCHES DIA."
1212 GOSUB9000
1220 NEXTTT
1230 PRINT:PRINT"WRITE DOWN ELEMENT LENGTHS APPEARING ON THE SCREEN AND ADD THE"
1235 PRINT"FOLLOWING CORRECTIONS TO THE OUTER MOST ELEMENT, FOR FINAL LENGTH."
1250 PRINT
1260 PRINT"ON 20 METERS ADD ";BC+0.66;"INCHES"
1270 PRINT"ON 15 METERS ADD ";BC+0.44;"INCHES"
1280 PRINT"ON 10 METERS ADD ";BC+0.24;"INCHES"
1285 GOSUB10000
1296 END
2000 DIMA(13),B(13)
2005 FORH=0TO13
2010 A(H)=0:B(H)=0
2020 NEXTH:REM CLEAR A AND B ARRAYS
2030 A(0)=0.49366:A(1)=0.49801:A(2)=0.49693:A(3)=0.49994
2035 A(4)=0.49528:A(5)=0.4705:A(6)=0.48963:A(7)=0.48366:A(8)=0.4804
2040 A(9)=0.48028:A(10)=0.469:A(11)=0.46798:A(12)=0.45232:A(13)=0.44811
2050 DIMR(6),DE(6),DR(6)
2060 FORQ=0TO6
2070 R(Q)=0:DE(Q)=0:DR(Q)=0
2080 NEXTQ:REM CLEAR R, DE, DR
2090 IFAD=0.875THEN2240
2100 RA=0.875/(2*RL)
2110 RB=AD/(2*RL)
2120 KA=1/RA
2130 KB=1/RB
2140 FORJ=0TO13
2150 F1=(1-((10.7575*(LOG(KA)/LOG(10)))-8)↑-1)/(2*A(J))
2160 XX=((215.15*(LOG(KA)/LOG(10)))-160)*((1/F1)-F1)
2170 AA=XX/((215.15*(LOG(KB)/LOG(10)))-160)
2180 F2=(-AA+((AA↑2)+4)↑0.5)/2
2190 B(J)=(1-((10.7575*(LOG(KB)/LOG(10)))-8)↑-1)/(2*F2)
2200 NEXTJ
2210 R(2)=B(0):R(3)=B(1):R(4)=B(2):R(5)=B(3):R(6)=B(4):DE(2)=B(5):DE(3)=B(6)
2215 DE(4)=B(7):DE(5)=B(8):DE(6)=B(9)
2220 DR(3)=B(10):DR(4)=B(11):DR(5)=B(12):DR(6)=B(13)
2230 GOTO 2260
2240 R(2)=A(0):R(3)=A(1):R(4)=A(2):R(5)=A(3):R(6)=A(4):DE(2)=A(5):DE(3)=A(6)
2245 DE(4)=A(7):DE(5)=A(8):DE(6)=A(9)
2250 DR(3)=A(10):DR(4)=A(11):DR(5)=A(12):DR(6)=A(13)
2260 RS=(RL*R(P))/2:REM REFLECTOR NORMALIZED 1/2 LENGTH
2270 DE=(RL*DE(P))/2:REM DRIVEN ELEMENT NORMALIZED 1/2 LENGTH
2280 DR=(RL*DR(P))/2:REM DIRECTORS NORMALIZED 1/2 LENGTH
2290 RETURN
3000 PRINT"{CLR}"
3010 PRINTP;"ELEMENT BEAM AT";CDF;"MHZ"
3020 PRINT
3030 PRINTRS;"INCHES=NORMALIZED REFLECTOR 1/2 LENGTH"
3040 PRINTDE;"INCHES=NORMALIZED DRIVEN ELEMENT 1/2 LENGTH"
3050 PRINTDR;"INCHES=NORMALIZED DIRECTORS 1/2 LENGTH
3060 RETURN
4000 PRINT:PRINT"PLEASE WAIT I'M THINKING"
4010 RETURN
5000 PRINT"{CLR}":REM CLEAR SCREEN
5010 PRINTP;"ELEMENT BEAM AT";CDF;"MHZ"
5020 FORAB=0TO(N-1)
5030 PRINTL(AB);"INCHES LONG",D(AB);"INCHES DIA."
5040 NEXTAB
5050 RETURN
5090 IFZ$="N"THEN6090
6000 OPEN1,4,0
6005 PRINT#1:PRINT#1:PRINT#1:PRINT#1:PRINT#1
6010 PRINT#1,"MONO BAND ANTENNA DESIGN":PRINT#1
6020 PRINT#1,"CENTRAL DESIGN FREQUENCY";CDF;"MHZ":PRINT#1
6030 PRINT#1,P;"ANTENNA ELEMENTS":PRINT#1
6040 PRINT#1,BD;"INCHES BOOM DIAMETER":PRINT#1
6050 PRINT#1,N;"ANTENNA SEGMENTS":PRINT#1
6060 PRINT#1,T;"INCHES TOTAL ORIGINAL LENGTH":PRINT#1
6070 PRINT#1,S;"INCHES TOTAL EQUIVALENT LENGTH":PRINT#1
6080 CLOSE1
6085 GOSUB8000
6090 RETURN
7040 RETURN
8000 IFZ$="N"THEN8060
8010 OPEN1,4,0
8020 PRINT#1,RS;"INCHES=NORMALIZED REFLECTOR 1/2 LENGTH"
8030 PRINT#1,DE;"INCHES=NORMALIZED DRIVEN ELEMENT 1/2 LENGTH"
8040 PRINT#1,DR;"INCHES=NORMALIZED DIRECTORS 1/2 LENGTH":PRINT#1
8050 CLOSE1
8060 RETURN
9000 IFZ$="N"THEN9040
9010 OPEN1,4,0
9020 PRINT#1,L(TT);"INCHES LONG";D(TT);"INCHES DIA."
9030 CLOSE1
9040 RETURN
10000 IFZ$="N"THEN10070
10010 OPEN1,4,0
10015 PRINT#1
10020 PRINT#1,"ADD CORRECTIONS TO OUTER MOST ELEMENTS, FOR FINAL LENGTH"
10030 PRINT#1:PRINT#1,"ON 20 METERS ADD";BC+0.66;"INCHES"
10040 PRINT#1,"ON 15 METERS ADD";BC+0.44;"INCHES"
10050 PRINT#1,"ON 10 METERS ADD";BC+0.24;"INCHES"
10055 PRINT#1:PRINT#1:PRINT#1:PRINT#1:PRINT#1
10060 CLOSE1
10070 RETURN

READY.
```

is calculated in **lines 2260-2280. Lines 770-970** calculate the resonant half-length of the element you have created by the lengths and diameters you entered earlier. **Lines 972-974** and **5090-10070** print a hard copy if you answer "Y" to the hard copy prompt.

You'll see printed on the screen the type of Yagi you have selected, the resonant half length each element should be (the "normalized" length) the mechanical length/diameter of each segment, and the actual electrical half-length of the element you've created. The program will now allow you to change individual segment lengths until your element's electrical half length equals the normalized half-length. You then add a small correction to the outermost segment (due to boom diameter and boom to element clamping system) to get the final element half-length.

W2PV broke the center of the driven element and used a balanced feed. I have used the gamma match successfully.

results

Three monoband Yagis have been constructed

design example 1: 3-element Yagi
f_0: 14.175 MHz
boom: 2 inches

element	length (inches)
normalized reflector	207.495860
normalized driven	204.066957
normalized director	195.621141

segment specifications

diameter (inches)	length (inches)
1.25	36
1.125	50
0.875	37
0.75	20
0.625	44
0.5	16
0.375	12.75

total original length
= 215.75 inches
total equivalent length
= 207.51345 inches

design example 2: 6-element Yagi
f_0: 14.175 MHz
boom: 2 inches

element	length (inches)
normalized reflector	206.378917
normalized driven	200.239937
normalized director	187.061662

segment specifications

diameter (inches)	length (inches)
1.25	36
1.125	50
0.875	37
0.75	20
0.625	44
0.5	16
0.375	11.5

total original length
= 214.5 inches
total equivalent length
= 206.397459 inches

Note 1: In each example, element has seven segments and last dimension is length before final correction due to boom clamping.

Note 2: If you cannot achieve the above numbers out to at least five places to the right of the decimal point, especially the normalized length and the total equivalent length, go back and check your program. Close is not good enough. Small inaccuracies are multiplied many times as segment lengths change.

using the program. When mounted on towers, not one resonated more than 4 kHz from the central design frequency. But the only way to really test an antenna is in actual combat — i.e., under contest conditions: the CQWW phone contest yielded 1.68 million points in 1983 and 2.43 million points in 1984. From 1983 to 1984, nothing changed except the antennas; all primary Yagis were homebrewed, using this program, and the secondary antennas were commercial beams redimensioned from the program.

W2PV hoped that others would build his superior beams and report the results. I've built them and am pleased to report that their performance is far better than any commercial antenna I've ever used from my QTH.

acknowledgement

Because this was my first attempt at programming, my sincere thanks are due to my XYL, N3DPB, who gave a few pointers on programming, and to WA3HQX, who converted my original program to the Commodore format. The program has been converted into Apple, Atari, IBM PC, and Radio Shack Color formats. Copies of these listings are available from me for one dollar (copying cost) plus a business-size SASE with 39 cents postage.

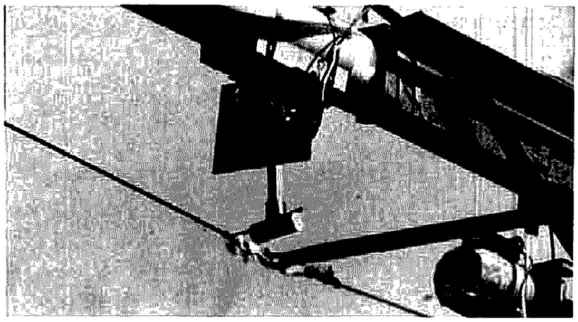
reference

1. James J. Lawson, W2PV, "Yagi Antenna Design," *ham radio*, January, February, May, June, July, September, October, November, December, 1980. (Back issues are available from *ham radio* at $3 each. Contact Ham Radio's Bookstore, Greenville, New Hampshire 03048.)

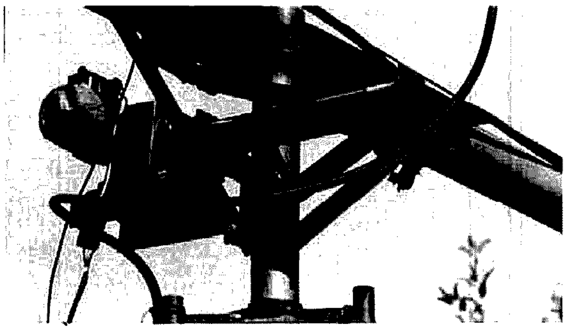
ham radio

the weekender

an inexpensive elevation indicator



Simple method of converting elevation angle to a voltage. Note weather-resistant housing.



Elevation drive: reversible gear motor turns leadscrew of 3/4-inch all-thread.

If you have a high input impedance (≥1 Megohm) multimeter — preferably digital — and a single-turn, precision, linear taper potentiometer, then you have everything you need to make an inexpensive, accurate elevation indicator for OSCAR EME applications.

After discarding my store-bought elevation rotator in favor of a heavy-duty home-brew model, it became necessary to provide some means to determine where the antenna was pointing without having to "eye-ball" the moon and the antenna. This resulted in the construction of the indicator shown in **photos A** and **B**.

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The digital multimeter is used to read 0 to 0.9 volt, corresponding to 0 to 90 degrees in elevation. R2, a precision potentiometer, is used as a voltage divider, with the indicating voltage read off the moving arm of the potentiometer. It is mounted on a piece of heavy-duty circuit board, of any convenient size with three-wire cable, preferably shielded, running from the antenna to the indicating unit in the ham shack. A suitable bracket for mounting it to your antenna may be soldered to the PC board.

Assuming your potentiometer is in the 5 to 10 kilohm range, required current will be less than 1 mA. Ordinary zinc-carbon cells are quite satisfactory for this purpose; in fact, they will probably die of old age if switch S1 is used. Voltage regulation is not necessary. R1, used to drop the voltage to approximately 3.6 volts, is connected to terminal 3 of the precision potentiometer for calibration purposes as the batteries gradually discharge (see **fig. 1**).

construction

Because most precision potentiometers don't have threaded bushings around the shaft output, it will probably be necessary to use a good adhesive (Super-Glue,® contact cement or epoxy) or a clamp to mount the potentiometer to the PC board. A 6-inch (15 cm) length of 3/8-inch (0.95 cm) square key-stock is recommended for the pendulum shaft. A 1/4-inch (0.64 cm) hole (assuming the potentiometer shaft is 1/4 inch) is drilled through the key stock, about 1/2 inch (1.3 cm) from the top end. Immediately above this, and at a right angle to the 1/4-inch (0.64 cm) hole, drill a hole to accommodate a No. 8 machine screw. Then cut a hacksaw slit from the top end of the key-stock into the 1/4-inch (0.64 cm) hole so as to provide a clamp around the potentiometer shaft when the No. 8 machine screw is inserted and tightened with a nut.

The pendulum weight may be made of any convenient material that will provide sufficient weight without too much bulk. I used a 3/4-inch plumbing-type copper T-fitting filled with lead (see **fig. 2**). Its estimated weight is about 10 ounces (283 grams).

As shown in **photo A** the potentiometer is enclosed for weather protection. The enclosure is made of very thin (about 0.010 inch) (0.25 mm) double-sided PC board, soldered in place. The opening is so oriented that protection is provided when the antenna is elevated. Similar protection on the shaft-pendulum side of the PC board is advisable, although I chose not to take this step. I did put some silicone grease in this opening, however, and even after one year had no problems, despite the occurrence of a severe storm with baseball-size hail. An automotive trailer light con-

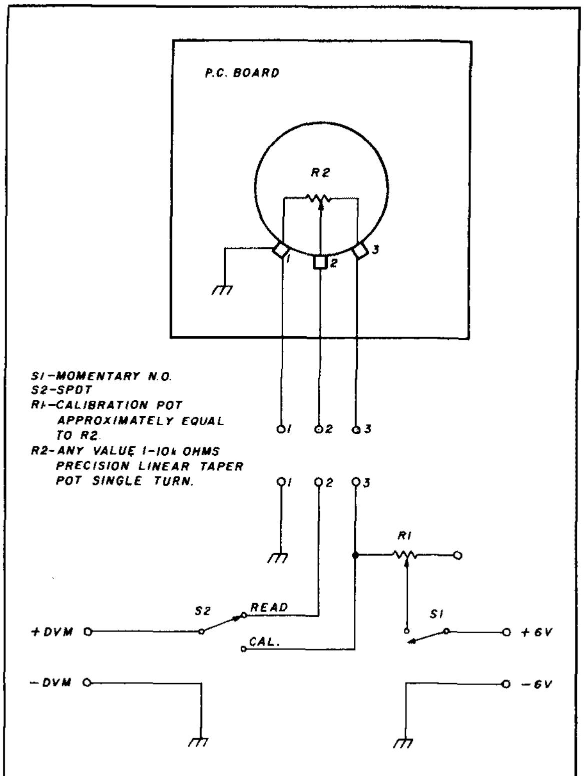

fig. 1. Elevation indicator wiring diagram.

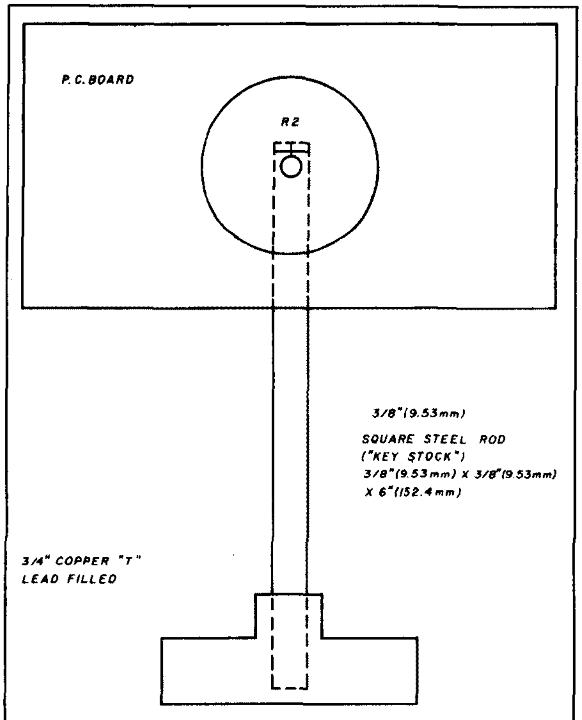

fig. 2. Construction details, R2 assembly.

nector is used for the wiring (as a quick make/break interface).

The read-out portion, shown in the bottom part of **fig. 1** was built in a small box, fabricated from circuit board, large enough to hold the batteries, switches and resistor R1. Although I used four D cells, I think smaller batteries, such as size AA, should be satisfactory. Since it is only occasionally necessary to monitor elevation visually, I would suggest that S1 be a normally open momentary pushbutton switch to be depressed only when readout is desired. S2 is a single pole, double throw switch. It is normally left in the "Read" position and is connected to the lead going to the variable arm of R2. Suitable pin plugs are provided for the new leads at the +DVM and −DVM points.

calibration

All the precision potentiometers that I'm familiar with permit continuous rotation, do not have mechanical stops, and offer almost 360 degrees (of rotation) of usable variable resistance. The potentiometer should be oriented on the PC board (or the pendulum adjusted) so that the arm terminal of the potentiometer, reads "0" voltage when the bottom edge of the PC board is horizontal and increases progressively to a reading of 0.900 volt when the bottom edge is vertical. A carpenter's level, with the bubble carefully centered during adjustments, with the circuit board holding the potentiometer and the pendulum in a vice, or otherwise strongly secured, will provide sufficient accuracy.

The slightest upward movement of the antenna should result in a voltage indication of 0.001. After the zero point has been established, things can be permanently secured. Using the carpenter's level, move the assembly so that the bottom edge of the circuit board is vertical. With S2 in the "Read" position, adjust R1 to give a reading of 0.900 volt, which corresponds to an elevation of 90.0 degrees. The unit is now calibrated and will indicate changes of 0.1 degree, to an accuracy of 0.5 degree. When the potentiometer pendulum assembly is mounted on the antenna, it is necessary only to level the bottom edge in the zero position, with the antenna horizontal, in order to maintain calibration. After the 90.0-degree position has been established and initially calibrated, put S2 in "Calibrate" position and note the indicated voltage. It is advisable to record this for later reference. You are now assured that you will be in calibration when R1 is adjusted to give this reading, while in the "Calibrate" position. For maximum accuracy, the tower must be perfectly vertical, and the vertical rotational axis perfectly horizontal.

conclusion

The author makes no claim of originality in this type of elevation readout, although I have seen nothing as simple as this one. Most others use three-turn potentiometers with 10.8 volts, with readouts taken from some midrange, such as 1.000 to 1.900 volts, and some means of removing the first digit. The readout described here is simple, inexpensive, and accurate. Invariably, with my EME array adjusted according to a computer-predicted position, I can sight down the boom of one of the antennas and know its direction will be "rifle scope" accurate.

Caution: I assume, but do not know, that all DMMs have a high input resistance, so as not to unduly load the portion of R2 between the arm and ground. If your DVM does not have a high resistance input, do not use it.

ham radio

automatic temperature control

Prevent premature equipment failures by sensing and acting on heat changes

A few months back I acquired a solid-state VHF amplifier capable of running 200 watts output with only 3.5 watts of drive from my handheld. Because I do most of my VHF operating at home, I decided to put the amplifier in the garage with the rest of the ham shack. This facilitated connections to the power supply and antenna, and since the garage is connected to the kitchen of our house, a 25-foot "umbilical cord" of RG8-X coax allows me to wander about the kitchen and living room while continuing to QSO on 2 meters with 200 watts out.

The only problem with this arrangement is that a 200-watt amplifier draws a great deal of current and gets very hot, especially during the summer in our garage. A large muffin fan mounted at the back of the amplifier cooled it to a reasonable level, but I still had to go out to the garage to turn the fan on every time I wanted to use the amplifier. Although this procedure worked fine, it was an annoying task.

One day some DX suddenly appeared on 146.52. Without thinking, I grabbed my handie talkie and proceeded to make a 1200-mile contact on direct! Fantastic!

Yes, it was . . . until the transistors in the amplifier unsoldered themselves. Fortunately, the transistors themselves weren't ruined — I don't know why — and I was able to repair the damage. But I obviously needed a better method of controlling the temperature of the amplifier without having to remember to turn on the fan and without having to waste electricity by leaving the fan running all the time.

The thermostatically controlled AC outlet box described herein solved the problem. It features an adjustable thermostat built around a Motorola MC3423 overvoltage protection IC, a remote temperature sensor, and visual indicators of the state of the thermostat and the presence of AC voltage at the outlet. A flip of the front panel switch allows the thermostat to detect a falling temperature rather than a rising one, thus enabling the box to be used to turn on a heater or crystal oven during the winter months.

theory of operation

Figure 1 is a schematic of the thermostatically controlled outlet box. The heart of the circuit is U1, an MC3423. This IC was originally designed to function as an overvoltage protection device for DC power supplies. Normally, R6 and R7 would be two resistors placed across the output of the power supply. If the power supply voltage exceeds a critical amount (determined by the values of R6 and R7), pin 8 of the MC3423 goes high (approximately 2 volts <VCC). Normally, pin 8 would be connected to the gate of an SCR that had been bridged across the terminals of the supply. Voltage on the gate would cause the SCR to conduct, shorting the power supply and blowing a fuse or tripping a circuit breaker.

However, this very versatile IC can function as a thermostat by using an NTC (negative temperature coefficient) thermistor for R7. Using a potentiometer for R6 allows the thermostat circuit to be adjusted. When the power supply's voltage is constant, the resistance of R7 changes with temperature and "fools" the MC3423 into sensing an overvoltage condition, sending voltage to pin 8. As R7 warms up, its resistance falls. A greater value for R6 means a lower temperature will trigger U1 on, while lowering the value of R6 raises the temperature at which U1 triggers.

Substituting a PTC (positive temperature coefficient) thermistor for R7 — that is, a thermistor whose resistance decreases as its temperature decreases — would allow one to sense falling temperatures instead of rising ones. However, PTC thermistors are difficult

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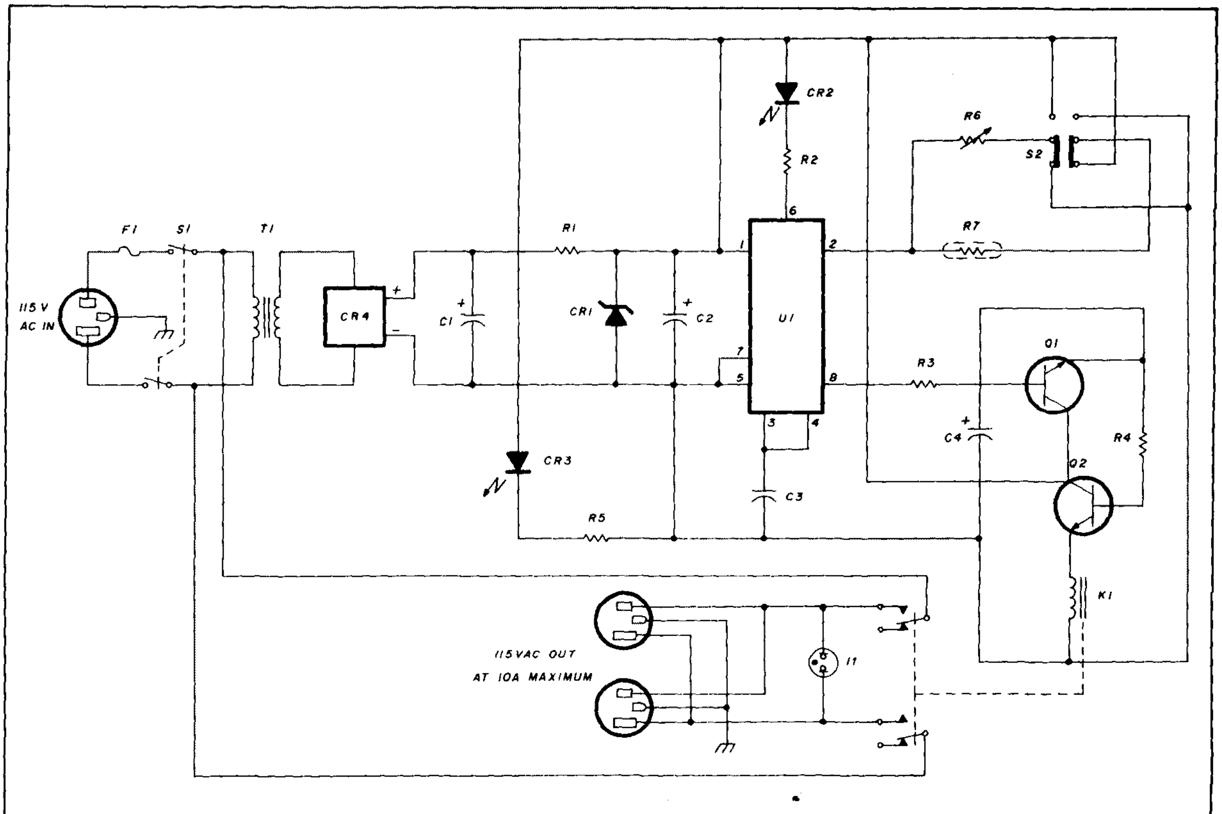

fig. 1. Schematic diagram of the thermostatically controlled outlet box.

item	description
C1	3000 μF, 35-volt electrolytic
C2	100 μF, 35-volt electrolytic
C3	0.1 μF disc ceramic
C4	330 μF, 35-volt electrolytic
CR1	13-volt, 5-watt zener diode (RCA SK-13 or equivalent)
CR2	amber LED
CR3	green LED
CR4	bridge rectifier, 50-volt, 1 ampere (RS 276-1161 or equivalent)
F1	10-ampere fuse
I1	miniature red neon lamp assembly (RS 272-708 or equivalent)
K1	DPDT relay, 10-ampere contacts at 125 VAC, 12-volt coil (RS 275-218 or equivalent)
Q1,Q2	2N2222 transistors or equivalent
R1	75 ohms, 5 watts
R2	270 ohms, 1/2 watt
R3	10 kilohm, 1/4 watt
R4	15 kilohm, 1/4 watt
R5	270 ohms, 1/2 watt
R6	50 kilohm, panel mount potentiometer
R7	NTC thermistor, 1/4 watt, 10 kilohms at 70 degrees F (FR-1001A or equivalent)
S1	DPST toggle switch, 10-ampere contacts
S2	DPDT miniature slide switch
T1	12.6-volt, 300-mA transformer (RS 273-1385 or equivalent)
U1	MC3423 overvoltage protection IC

to find in my area, and they are more expensive than NTC thermistors. S2 provides a simple way around this problem. Flipping S2, a DPDT slide switch, reverses the order in which R6 and R7 are connected to the power supply. As R7's temperature falls its resistance increases, thus allowing U1 to trigger when the temperature of R7 falls to a point determined by R6.

When U1 triggers it sends voltage to the base of Q1, a 2N2222, which conducts and allows voltage to be applied to C4 and the base of Q2, another 2N2222. C4 charges and Q2 conducts, thus keying relay K1, which applies 115 VAC to the dual outlet. Anything plugged into the outlet, such as a fan or heater, then turns on. U1 also turns on CR2, an amber LED, through pin 6, giving a visual indication that the triggering temperature has been reached. When the temperature changes to the point at which R7 can no longer hold U1 in its triggered state, U1 removes voltage from pin 8 and turns off CR2. C4, however, continues to energize K1 for approximately 1 minute, which prevents constant cycling of the relay and the devices connected to the dual outlet. Changing the value of C4 results in a fairly linear change in relay hold-in time for K1. For example, doubling the listed value of C4 gives a hold-in time of approximately 2 minutes,

while halving the value of C4 gives a hold-in time of approximately 30 seconds. I1, a small red neon indicator across the contacts of K1, provides a visual indication that the outlet is "live," even when CR2 indicates that U1 is not in its triggered state.

Transformer T1 and CR4, C1, CR1 and C2 provide a regulated operating voltage for the thermostat. A regulated voltage is not really necessary for this circuit, and CR1 and C2 could be omitted with little performance degradation. However, since U1 is really a voltage-sensing device rather than a temperature-sensing device, a regulated voltage across R6 and R7 insures that changing line voltages will not affect the temperature at which U1 triggers.

Capacitor C3 prevents line transients from triggering U1 by setting the IC's internal delay time. Increasing the value of C3 increases the time during which an "overvoltage" condition can exist before U1 triggers. If this thermostat is to be operated in an environment where transients can be picked up by the internal or external wiring, such as near a transmitter, the value of C3 may have to be increased to prevent false triggering of U1. Since the duration of transients will vary from location to location, you will have to determine the best value for C3 yourself. I have had to go as high as 0.5 μF in some locations, such as near a 2-kW homebrew amplifier.

Relay K1 has contacts rated at 125 VAC, 10 amperes maximum. Thus, F1 is a 10-ampere fuse intended primarily to protect K1 from damage if the load at the dual outlet becomes excessive. Of course, any relay can be used at K1, but using a relay with a lower coil impedance than 160 ohms may necessitate using a heftier transistor at Q2. Any relay you use, however, must be able to be energized by the voltage across T1. A relay whose contacts can handle more current may be necessary, in addition to a heftier switch at S1, if the thermostat is to be used to control some devices, such as space heaters, that draw large amounts of current when first turned on. The value of F1 would also have to be increased in such an instance. In the interests of safety, however, *do not, under any circumstances, omit F1*, and be certain that it is placed on the "hot" side of the AC line, not on the "neutral" side.

construction

Construction of the thermostat and outlet box is fairly straightforward, and layout is entirely noncritical. All the components, including T1 and K1, will fit on one small circuit board, with the exception of R7. This is covered with heat-shrink tubing and attached to the end of a piece of RG-174U miniature coax for remote sensing. R6, CR1, CR2, I1, S1, S2, and the AC outlet are mounted on the front panel. **Figure 2** is a

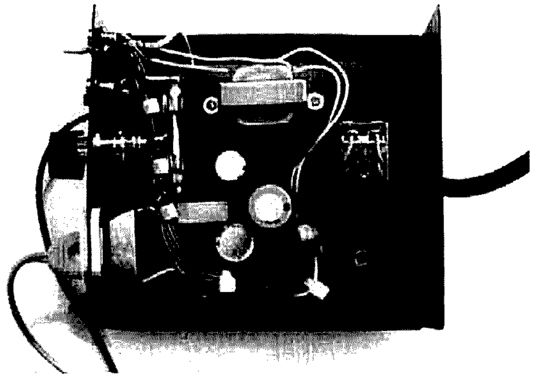

fig. 2. Internal view, with top cover removed. F1, T1, and K1 are mounted near the top of the board. U2 is barely visible beneath the wiring at the lower left corner of the board.

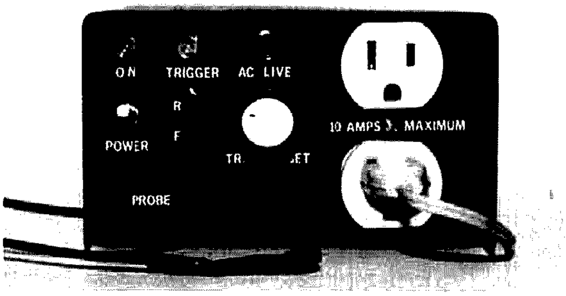

fig. 3. View of the front panel. Coiled black wire contains thermistor R7 (covered with heatshrink tubing) at one end.

photograph of the completed outlet box with the cover removed.

I suggest using sockets for CR4 and U1 to make replacement, if ever needed, easier. In addition, you should use at least No. 12 wire from the AC line to K1 and from K1 to the dual outlet.

operation

Using the thermostat box is easy. Just plug it into a 115 VAC outlet and supply power through S1 (see fig. 3). The green LED should come on, and the amber LED and red neon indicator may come on at this time. If they do, adjust R6 until the amber LED goes out; a minute later I1, the red neon indicator, should go out as C4 discharges and K1 opens. *Once again, remember that whenever I1 is on, 115 VAC is present at the dual outlet, and I1 provides a visual warning of the presence of potentially lethal voltage.*

Now set S2 to the condition you wish to monitor – that is, either rising or falling temperature. Place R7 at a convenient point where it will be exposed to the temperature you wish to control. For example, the sensor can be placed between the fins of an amplifier's

heatsink. Plug a fan into the outlet on the front panel, and adjust R6 until CR2 and I1 come on when the amplifier becomes hot enough to need cooling. As the fan blows cool air across the amplifier's heatsink, R7 will cool off and CR2 will go out. The fan, however, will continue to run for one minute.

That's all there is to it. As the amplifier heats up, the thermostat will turn on the fan to cool things down. When the amplifier is not is use, the fan will remain off.

other uses

This device can be used anywhere around the shack or the house where you wish to sense a rising or falling temperature. For example, it can function as a freezer alarm by plugging a 115 VAC alarm bell into the outlet and adjusting the thermostat to trigger whenever the temperature rises above 32 degrees F. It can be used to turn on a crystal oven when the temperature inside an oscillator falls below 80 degrees F (27°C). It can even be used to turn on a floor fan or space heater whenever the temperature in your shack rises or falls to uncomfortable levels. Anywhere you need to monitor a rising or falling temperature and turn on some device at a critical point, this thermostat will come in handy.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

protecting equipment

Spring is typically the season in which most VHF/UHFers dust off their gear and get ready for all those good tropospheric and sporadic-E openings. It's also the time of year when static electricity and lightning tend to increase. Between the increased operating activity and the forces of nature, equipment failures can and do happen! Whether the result of carelessness or the ravages of nature, damage to expensive equipment can be frustrating. But with proper know-how and a little bit of effort, failures can be anticipated and prevented.

Many types of failures can affect Amateur equipment. — in-band RF, out-of-band RF, low frequency and supply transients, the forces of nature, damage in physical handling, and component overstressing are just a few.

in-band RF

Probably the prime source of failure in Amateur VHF/UHF equipment, in-band RF most often affects receivers. Simply put, the receiver input stage is subjected to RF on the frequency of interest at a level that exceeds the breakdown of the first active device. With the proliferation of solid-stage receivers over the last decade — especially on 70 cm and above, where high transmitter power (1000-1500 watts output) is required, — this phenomenon is now quite common. The primary reasons for this type of failure are inadequate T/R (transmit/receiver) relay isolation and improperly timed relay sequencing.

relay isolation

Most Amateurs use T/R relays that are either the bladed or the lever arm type. The bladed type (**fig. 1A**) is the most common and is similar to the everyday low-frequency relay except in that a moving arm is built into a coaxial airline structure. The contacts are capacitance junctions and the spacing between the moving contact and the input or output connector contact determine the isolation between the transmitter and receiver. There are many tradeoffs with this type of relay because the spacing and the size of the contacts determine the isolation, VSWR, and power handling capability of the relay.

If the capacitance across the relay junction is known, the isolation can be calculated using the following formula:

$$\text{isolation} = 10 \log_{10} [1 + (X_c/2Z_0)^2] \qquad (1)$$

where isolation is in dB, X_c is the capacitive reactance of the junction, and Z_0 is the transmission line impedence, typically 50 ohms. For example, a typical relay junction has a 0.1 pF capacitance. Therefore, at 500 MHz the capacitive reactance is 3180 ohms and the isolation will be approximately 30 dB. (For those who do not want to work out the mathematics, **fig. 2** shows isolation versus capacitance values.) Note that the isolation across a purely capacitance junction decreases approximately 6 dB every time the frequency is doubled. Consequently, a relay with 40 dB of isolation at 2 meters will have only approximately 30.5 dB of isolation at 70 cm — quite a decrease!

Some relay designers add a set of grounding contacts across the open circuited contacts (**fig. 1B**). This can significantly increase the isolation. However, this places a short circuit across the attached device, typically a low-noise preamplifier. This may cause transients when switching or oscillations while in the transmit mode.[1]

The lever arm relay configuration (**fig. 1C**) is more complex and hence more costly, but has higher isolation because this construction increases the spacings between the input and open circuited port of the relay. However, it is likely to have a lower power handling capability than the bladed type.

The relays most often used by Amateurs (and often found at flea markets) include the Amphenol 300, Dow-Key DK-60, M/A-Com 7524, and the Transco "Y." Typical isolation versus frequency is shown in **fig. 3** for some of the types mentioned. A typical solid-state receiver will withstand 10 milliwatts or +10 dBm (dB above a milliwatt) at its input without damage. Therefore, at an RF power level of 1500 watts (+62 dBm), a relay with at least 52 dB of isolation should be used.

The Transco type-Y relays I've pur-

chased at flea markets have shown high insertion loss on one or both of the paths. This is easily checked with an ohmmeter. When the contacts are energized, the resistance should be less than 1 ohm. Try cycling the relay several times to check for intermittencies, a common occurrence.

If the resistance is high, unscrew the appropriate connectors with a narrow-width open end wrench (if the relay has set screws on the connectors, remove them first). Then burnish the contact point on the end of the connectors. Next, reach inside the center compartment and burnish the complementary contacts. (Better yet, remove all three connectors, carefully noting the position from which each was taken.) Clean all internal contacts and connectors and as a final measure use solvent or flux remover on all contacts to eliminate any film. Then rotate the connectors 120 degrees so that a different connector is placed in each position; this will increase the likelihood that each connector contact will be at a new and clean point. Retest the contact resistance. If you have a good RF test setup, measure the insertion loss. It should be less than 0.2 dB through 450 MHz. Isolation is seldom a problem.

All the relays just metioned are available with type "N" connectors, the preferred connector type for 2 meters and higher frequencies. However, this connector type is power-limited to approximately 500 watts at 500 MHz and commensurately lower at higher frequencies. I know that many Amateurs are exceeding this power level on "N" connectors; if you're one of them, keep your VSWR low and never try to "hot switch" RF power.

relay configurations

Some of the relays mentioned will not provide the isolation necessary to protect receivers, especially on the higher frequency bands. Therefore, a second relay is often cascaded with the T/R relay (see **fig. 4A**) to further increase receiver protection.[1]

Because this second relay is switching low power, it doesn't have to be a high power type. Often relays with BNC connectors are used at the lower frequencies and SMA connectors at UHF. Furthermore, this relay doesn't have to have high isolation since the primary T/R relay is capable of providing most of the needed attenuation.

If two relays with similar isolation are connected back-to-back with the shortest possible connection, the combined isolation is only 6 dB greater than the single relay isolation.[1] If the same two relays are placed an electrical quarter wavelength apart at the operating frequency, the isolation will

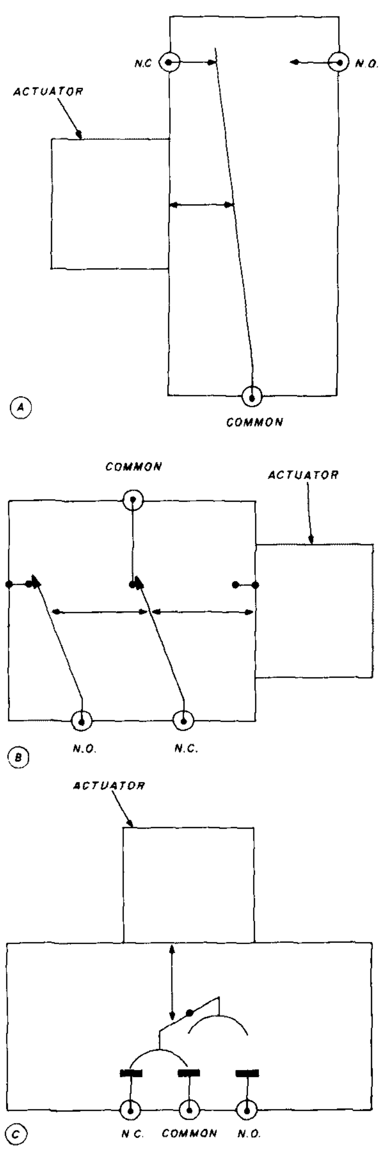

fig. 1. Common types of coaxial relay switches: *(A)* bladed, *(B)* bladed with additional shorting contacts, and *(C)* lever arm.

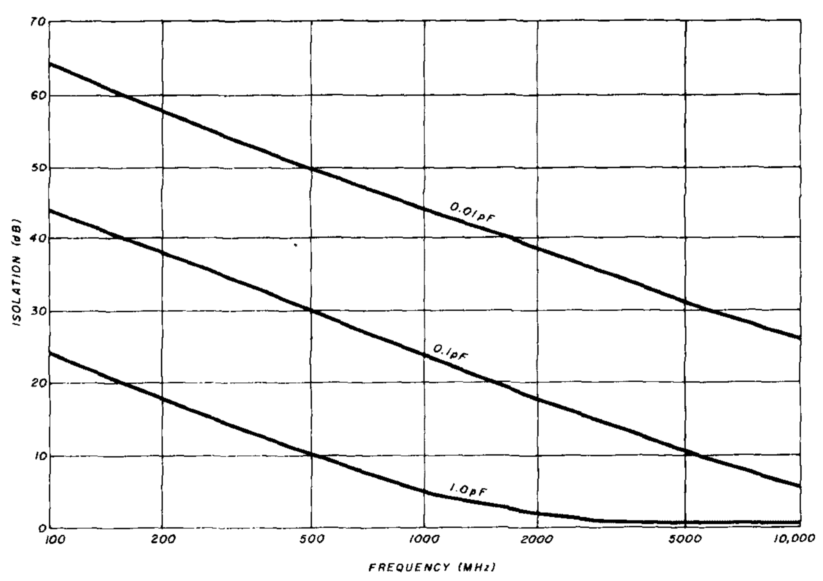

fig. 2. Isolation versus frequency when a capacitor is connected in series with a 50-ohm transmission line per *eq. 1*. Other capacitance values can be interpolated.

be approximately 6 dB greater than the combined relay isolation. For example, if two relays each with 30 dB of isolation are connected is cascade, the theoretical maximum isolation will be 66 dB.

*Never use one-half wavelength spacing between T/R relays since the attenuation can theoretically go to 0 dB (**fig. 5**).*[2] *For practical and theoretical considerations, one-tenth wavelength spacing is all that is recommended because it will keep losses at a minimum and decrease isolation by about only 6 dB from the maximum possible!*

Figure 5, an updated version of the one shown in reference 1, will yield typical isolation values at a glance for a 30-dB isolation relay (the example discussed at the beginning of this article).

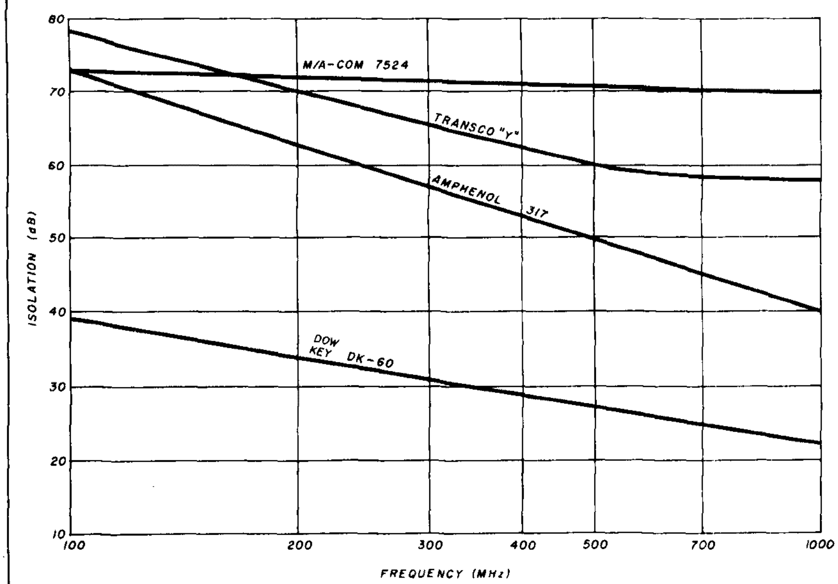

fig. 3. Typical values of isolation on some of the more popular T/R relays used by Amateurs.

Other values of isolation can be calculated. The relay capacitance must first be measured or calculated using **eq. 1**. Then use the following equations:

$$isolation = 10 \log_{10} 0.25 \; [4 + (2X_N \cos\theta - X_N^2 \sin\theta)] \quad \textbf{(2)}$$

where isolation is in dB, X_N is per **eq. 3** (below), and θ is the electrical spacing of the relay contacts in degrees.

$$X_N = \frac{Z_0}{2\pi fC} \quad \textbf{(3)}$$

where f is the frequency of operation in Hz, C is the open-circuited capacitance of the relay contacts in farads, and Z is the transmission line impedance in ohms.

remote preamplifiers

Remotely located or antenna-mounted preamps are becoming very popular, especially on 2 meters and up, on OSCAR 10, and on EME. The dual relay scheme illustrated in **fig. 4A** is highly recommended in these applications. Note that the preamplifier is terminated with a 50-ohm load when in the transmit mode, not a short or open circuit as discussed earlier in this article and in reference 1.

Finally, all T/R relays should have adequate time to switch before any transmitter power is applied. This is easy to accomplish if a high-voltage

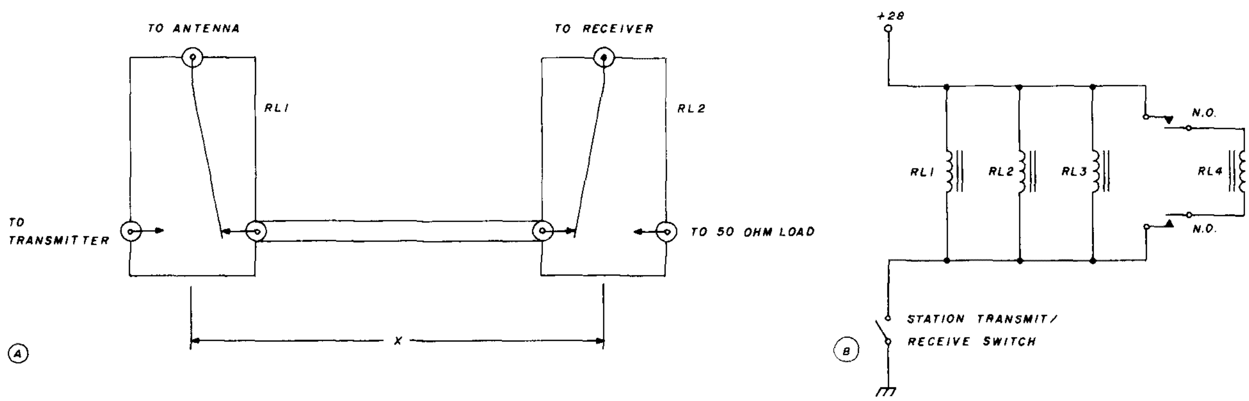

fig. 4. Recommended relay switching schemes. (*A*) see text for explanation of relays. Distance "X" should typically be 0.1 to 0.25 wavelength as explained in text. (*B*) RL3 is a typical DC type relay with double pole contacts. RL4 is the high-voltage relay in the transmitter power supply. (See text for further explanation.)

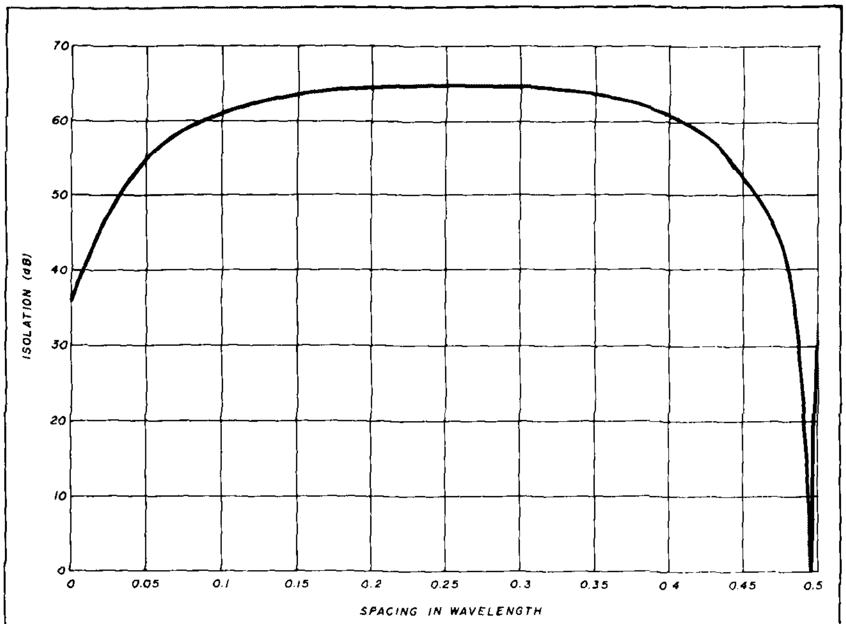

fig. 5. Additional isolation available from cascade-connected relays with a connecting transmission line.

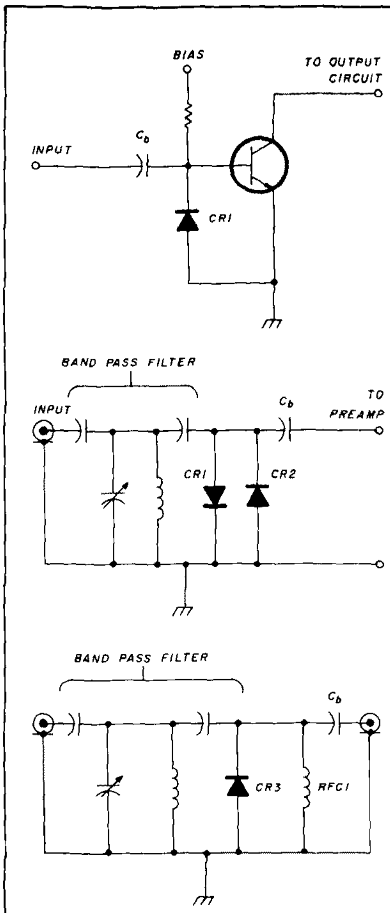

fig. 6. Typical recommended diode protection schemes for receivers. C_B is a blocking capacitor. CR1 and CR2 are hot-carrier types as described in text. CR3 is a PIN diode. RFC1 is an RF choke for DC return.

relay is used in the transmitter as discussed in February's column.[3] I've often seen complex circuitry that relies on R/C time constants and such. Timing capacitors, especially of the electrolytic type, are often unreliable and may decrease in capacitance as they age.

Figure 4B is a simple semi-foolproof way to add some time delay sequencing with the addition of only one low-cost DC relay. In this scheme, the extra relay will have to first close, which will add about 10 milliseconds delay before power is applied to the high voltage relay. When the T/R switch is opened, the high voltage relay loses power immediately and turns off.

One last caution is in order. Many modern transceivers, HF and VHF/UHF alike, can inadvertently transmit a short burst or pulse of RF either when they're first turned on or when they're switching modes. Depending on the station configuration, this could spell disaster to an inline preamplifier that is not properly protected.

input limiters

Relays are fine, but you may want some extra built-in protection. A simple low-power limiter was described in reference 4. A low-cost hot carrier diode is placed across the base to emitter junction of the bipolar transistor as shown in **fig. 6A**. This type of limiter will protect only up to about 1 watt. Hot carrier diodes used as limiters should have a low junction capacitance and a low forward resistance. For example, the M/A Com MA4882 or NEC ND4981-7E are recommended. The Hewlett Packard 5082-2810 and its equivalents are not recommended since the laboratory tests I conducted showed that they have a high resistance at increased current.

Often I see back-to-back diodes indiscriminately placed across the input of a receiver. Out-of-band signals — for example, FM, TV, and broadcast — can often induce enough voltage at a receiver input to drive these diodes into conduction and cause spurious signals to appear. Point contact diodes are the most susceptible, followed by hot-carrier and then silicon types. Therefore, if you use limiter/protection diodes at a receiver input, they should be located after a band-pass filter as shown in **fig. 6B**. Remember, hot carrier diodes can handle only up to about 1 watt of incident power.

Finally, commercial designers frequently use PIN diode limiters, especially above 500 MHz (see **fig. 6C**). These unique diodes turn on quickly, then drop to a very low resistance and reflect the incident RF. Most PIN diodes are able to handle much more power than hot carrier diodes. Most microwave diode suppliers manufacture these devices, but discussion of a recommended device and circuitry will have to wait for a future column.

out-of-band protection

The input protection schemes just mentioned are not always sufficient. Other RF problems can inadvertently occur from out-of-band signals. For instance, some years ago I advocated the use of an "idiot" diode in preamplifiers to prevent application of improperly polarized power supplies.[4] Eventually I placed idiot diodes into all my preamplifiers.

No preamplifier failure has ever occurred from improper supply polarity. However, soon after installing the idiot diodes, I began experiencing random burnout, especially on my J-FET preamplifiers, which are usually quite rugged. I tried all types of input filtering, without success. This random problem continued for over a year. Finally I realized that the burnout occurrences increased during the winter, a time of low electrical storm activity and the time of year when I spend many hours DXing on the low end of 80 meters.

With the help of a digital voltmeter, a second operator, and careful measurements, I found that the 80-meter RF was coupling to my 12-volt power supply leads going into my preamplifiers. The idiot diode was rectifying the RF and adding it to the 12 volts from the supply! In one case, the actual DC at the drain of a 2-meter preamplifier was 26 volts with the 80-meter kW in operation.[5] The reason this probably did not affect my bipolar preamplifiers was that they also had zener diode biasing, which limited the voltage across the device.[4]

The solution is simple. Place a 0.1 μF ceramic disc capacitor — remember, this is low frequency — on the DC line where it enters each preamplifier. (Larger capacitance values are not recommended because they may not act as a good RF bypass.) Any RF trying to enter the power line to the preamplifier is now bypassed and no longer seen by the idiot diode. The only failure I've experienced since that time was in a preamplifier that I forgot to bypass!

High-pass or bandpass filters are always recommended ahead of a receiver because they prevent RF, especially at HF, from entering the input of the preamplifier.[6,7] When selecting a filter, a capacitor input type is recommended per reference 8. Filters that use a shunt inductor at the input (such as a loop on a cavity filter) are not recommended because they usually have insufficient attenuation at lower frequencies.

low frequency and supply transients

Several types of failure modes are induced at low frequencies, especially in power supplies. One type is spikes or transients on the AC mains. Laboratory tests show that 500 to 1000 volt transients are always occurring, especially when motors or inductive devices are turned on and off. A shunt capacitor across the AC line helps, but is usually insufficient protection.

Zener diodes are often too slow to turn on and usually have limited power handling capacity. Modern protection devices such as the MOV (metal oxide-varistor) were specifically designed to operate at high frequencies and clamp incoming spikes similar to the action of a back-to-back zener diode. Such devices are inexpensive (less than $1.75) and readily available for 130 and 230-volt AC power. General Electric part No. V130LA10A or V250LA20A are suggested units for 130 and 250 volts AC, respectively. Radio Shack stocks similar MOVs. I highly recommend that you place one of these devices across the primary of all power supply transformers after the fuse as shown in **fig. 7A**.

General Semiconductor Industries, Inc. makes several products under the names TransZorb™ and ThyZorb.™ These devices, available from 5 to 700 volts, are particularly noted for their speed and power handling capability.[10] They can also be used across AC lines and wherever fast-acting zener diodes are desirable.

For many years I've been preaching that you should always use a dedicted power supply for all receiver circuits.[6] Using a preamplifier power supply for relays is not recommended. When relays are deenergized, inductive spikes are produced — and these can easily destroy a preamplifier if it's connected to the same supply. Even if you use a separate relay supply, it's best to place a 1N4004 or equivalent diode reversely polarized across all DC relay coils, T/R relays included (unless they're AC types) as shown in **fig. 7B**.

The manner in which solid-state devices are biased can also be a failure mechanism. The use of three terminal voltage regulators with protection diodes is highly recommended.[9] I strongly suggest the addition of a limiting resistor, however small (20 to 100 ohms is suggested), in series with each solid-state device to lower dissipation in case of runaway biasing. As another precaution, place a zener diode a few volts above the power supply voltage after the resistor. It will also serve as an idiot diode since most zeners have less than 1.0 volt drop when biased opposite to convention. These schemes are shown in **fig. 7C**.

AC line filters

Brute-force line filters are also recommended, especially to prevent damage from large surges typical of lightning strikes. A 50 to 100 μH inductor should be placed in series with the AC line where it comes into the ham shack and followed by an MOV or TransZorb™ of the appropriate rating as shown in **fig. 7D**. Note that the inductor must have an air core. Ferrite cores or rods will saturate and become ineffective if a large surge voltage should be induced onto the AC lines by a lightning strike.

the forces of nature

So far we've been talking mainly about protection against common man-made causes of burnout. But natural forces, in the form of static electricity or lightning, can be extremely destructive. Although statisticians assure me that even in the worst lightning-prone areas such as central

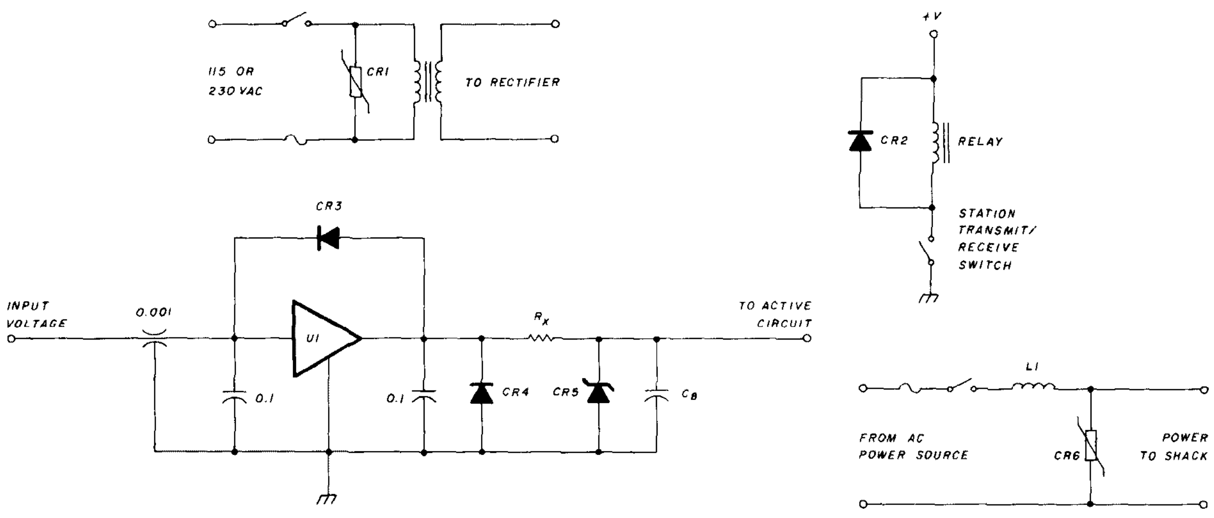

fig. 7. Recommended power supply protection schemes. C_B is a blocking capacitor. CR1 and CR6 are MOV type protection diodes. CR2, 3 and 4 are 1N4004 or equivalent (see text) and CR5 is a zener diode approximately 2 volts above the operating voltage of the circuit. R_X is typically 20 to 100 ohms (see text). L1 is a 50 to 100 μH RF choke made from 120 to 170 turns of No. 18 AWG or larger (depends on current to be drawn) enamel coated wire spaced wire diameter on a 1-inch (2.54 cm) diameter wooden dowel.

Florida, lightning strikes the same place only once every 10 years, *don't trust your luck.* My station has been hit more than once; the worst strike occurred in the Santa Clara Valley in California, a normally low probability area.

Because lightning protection has often been discussed in Amateur publications,[10] I will highlight only those problems that most affect VHF/UHFers. Experts tell me that when lightning strikes within 1000 feet (300-meters) of your equipment, damage will result. Even a 1-mile (1.6 km) separation from a lightning strike can result in damage caused by line voltage surges. Many simple techniques can be used to lessen or prevent lightning destruction to Amateur equipment; while some have been amply discussed in the literature, others have not, and I want to share them with you.

proper grounding is important

First, all antenna feeds should have a built-in ground return. This can often be incorporated as part of the feed system: for example, a "T" match or the balun connected to a metal boom or mast. The voltage breakdown of air at sea level with low humidity and room temperature between two flat surfaces is approximately 70 to 75 kilovolts per inch (28 to 30 kV/cm). A needle point, on the other hand, will break down at about 26 to 30 kV per inch (10-12 kV/cm). Therefore, a "Blitzbug"™ type of lightning arrestor installed in each transmission line will add an additional safety factor. Since this type of device has a point spacing of about 0.02 inch (0.05 mm), it should fire at 500 to 700 volts and handle reasonable follow-on current. The most common type of Blitzbug™ available is fitted with UHF connectors, but usable, with low VSWR, to about 450 MHz.

SVPs (surge voltage protectors) are becoming very popular. Basically they consist of a pair of electrodes properly spaced and hermetically sealed in a rare gas field. Breakdown voltages of 70 volts and up with low shunt capacitance (0.5 to 2 pF) are now available. Several manufacturers offer them for insertion in transmission lines. Remember that SVPs, while fairly quick to turn on, will not prevent damage from the large or sustained current typical of a direct hit as effectively as the Blitzbug.™

For best lightning protection it may be best to install both types of devices in cascade in your feedlines and spaced about one-tenth wavelength apart such as discussed in the relay section of this column. *In all cases, each of these devices should have a separate low-impedance ground return (see below).*

towers

Towers should never be higher than necessary. The highest object in a given local area, after all, is especially vulnerable to lightning strikes. Each time the height of a tower is doubled, the chance of its experiencing a direct hit increases by a factor of 4!

In the "good old days" lightning rods were installed on the rooftops of homes and tall buildings. I always thought they were meant to provide a discharge path, but some of my old-

timer friends tell me that they may in fact work in the opposite way, emitting electrons and thereby repelling or at least decreasing the probability of a direct strike! In this regard, the same experts tell me that a blunt or ball-shaped object would be a better way to ground a strike. I don't know which is correct. Maybe someone reading this column can enlighten us.

Regardless of the outcome of this question, all towers and antennas should be well grounded for lightning protection. A single ground rod is not sufficient. Two ground rods may be even less effective than expected because when they're positioned close together electrically, they couple to each other. Therefore, I recommend *at least* two ground rods spaced 10 to 15 feet (3 to 5 meters) apart, close to the base of a tower but at least 2 feet (60 cm) from any concrete used in the installation.

All ground rods should be at least 5 to 8 feet (1.5 to 2.5 meters) long, with a minimum diameter of 5/8 inch (16 mm). Always use high quality, well plated grounding rods (available from most electrical supply houses). Small low-cost ground rods often rust and become poor grounds after the first or second rain storm!

Grounding rods should be connected to the tower with a heavy (No. 6 AWG or larger) solid copper wire. This wire should be kept as straight as possible so that it forms a low impedance path for lightning (as just discussed). Keep all bends to a minimum. Also *keep ground return wires away from other wires or cables to minimize any coupling effects.*

A strong bonding connection should be made where the grounding wire connects to the tower and the rod. The inductance of the wire and its ability to handle the peak current of a direct lightning strike are very important. [I had a single aluminum 1/8 inch (3 mm) diameter ground wire that exploded open when hit by lightning. You should have seen the damage in the shack as the strike found other discharge paths!]

Lightning usually travels in a straight line. Therefore, whenever possible, bring all transmission lines away from your tower at right angles and in a reverse direction as shown in **fig. 8A**. This will cause any lightning to go directly to ground rather than down the transmission line, since the bend will act as a high impedance. In addition, you can loosely coil up to 3 to 5 turns of your feedline in a 4 to 8 inch (10 to 20 cm) diameter, depending upon the minimum bending radius, to act as an RF choke (see **fig. 8B**).

Remember that a direct hit can cause an extremely large current (20 to 100,000 amperes!) to flow. *Enclosing all transmission lines within an iron water pipe or tube (such as EMT) that is just a bit larger than the diameter of the line will considerably diminish the induced current. This is because of the limiting of the magnetic field by the steel (ferromagnetic material) tubing as shown in* **fig. 8C**. The preferred method is to use 20 feet (6 meters) of tubing for each line, but even a single 5-foot (1.5 meter) tube will help. If several shorter tubes are used, they need not be grounded or connected together. For economy's sake, several transmission lines can be placed inside a single tube. If this method is used and the tower is well grounded, the lightning will take the lowest impedance path — directly to the ground rods instead of through the ham shack!

When connecting transmission lines to a piece of gear, especially when rack mounted, bring the cables vertically from the floor level. Connect a suitable outside ground (such as the one just suggested for towers) to each shield at the floor level and at the base of the cabinet. This prevents the lightning from climbing vertically into the equipment.

lightning, static and RF

While a direct hit is hard to prevent and even more difficult to divert entirely, these suggestions will at least limit damage and perhaps prevent it entirely. Other measures can also help.

One of the best methods to protect equipment is to use a bandpass filter

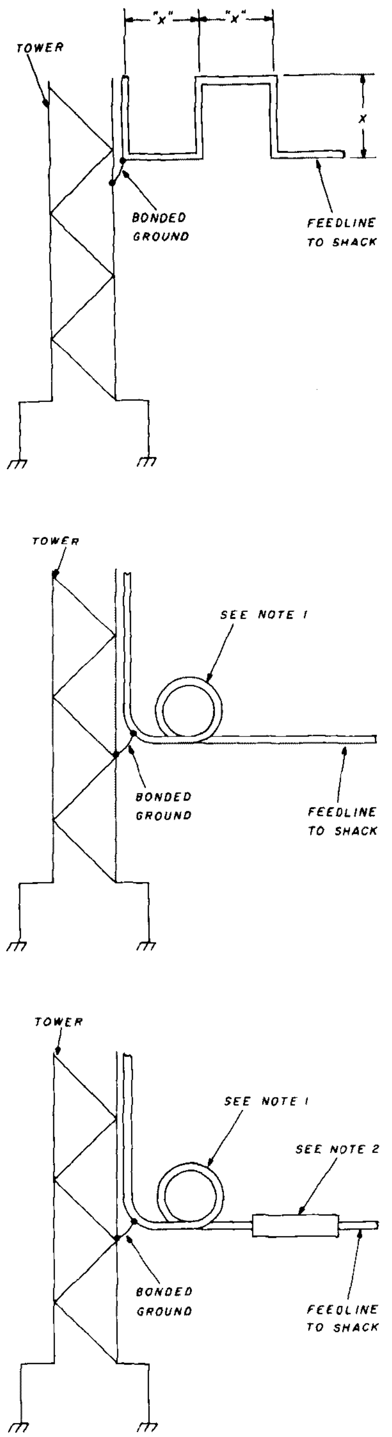

fig. 8. Typical methods for diverting lightning on transmission lines that are coming down a tower. "X" is typically 1 to 2 feet (30 to 60 cm). ***Note 1:*** **feedline wound to act like an RF choke (see text).** ***Note 2:*** **steel conduit (see text for size).**

(as described above). This is suggested because lightning is primarily a low-frequency phenomenon — i.e., below 30 MHz. A high-pass or band-pass filter (see **fig. 9A** with a high attenuation at lower frequencies will significantly decrease the energy entering a receiver. Unfortunately, this may not always be possible for those operating on the HF bands!

Likewise, keep all coupling capacitors at a minimum. A capacitance reactance of 2 to 5 ohms at the frequency of operation should be used. This would suggest the use of no more than 200 pF at 70 cm. When low noise transistors first became available for 70 cm operation, many preamplifiers were accidentally destroyed when connected to automatic noise figure meters that used a gas tube noise generator. This failure was caused by large coupling capacitors that responded to the low frequency transient emitted by the firing of the gas tube. When low value capacitors were used, the problems diminished considerably. Fortunately, modern automatic noise figure generators use solid-state diodes that don't have this problem.

The placing of a 5 to 10 kilohm resistor across your receiver inputs as shown in **fig. 9B** will also help, especially in elimination of static.[9] It's standard practice in the CATV industry to place 510,000-ohm resistors in shunt with drop cables (the ones that go to your home), presumably to bleed off any charge buildup.

Another technique is the use of a shorted quarter-wave stub. A coaxial "T" connector is placed in the transmission line and the stub is connected and grounded as shown in **fig. 9C**. This stub is a high impedance at the frequency of use. Therefore it has low loss, but is a DC short for any static or lightning.

physical handling

We may do all of the foregoing without paying attention to some of the basics such as using extreme care in handling and connecting equipment. All power supplies should be measured for proper voltages and should preferably be in operation when they are connected to the equipment. Some HF transceivers have external power supplies that put out high peak voltages when first turned on. Manufacturers recommend that the power supply should be turned on before the transceiver; this procedure should be reversed when ceasing operation. It is also recommended that whenever you connect a power supply, the ground return should be connected first. If not, the return current may flow through any coax cables back to the power supply and induce a spike or spark into the amplifiers. Transmission lines sometimes store a charge. Therefore, always discharge both sides of any incoming trasnmission lines to chassis ground before connecting them to the input of a receiver.

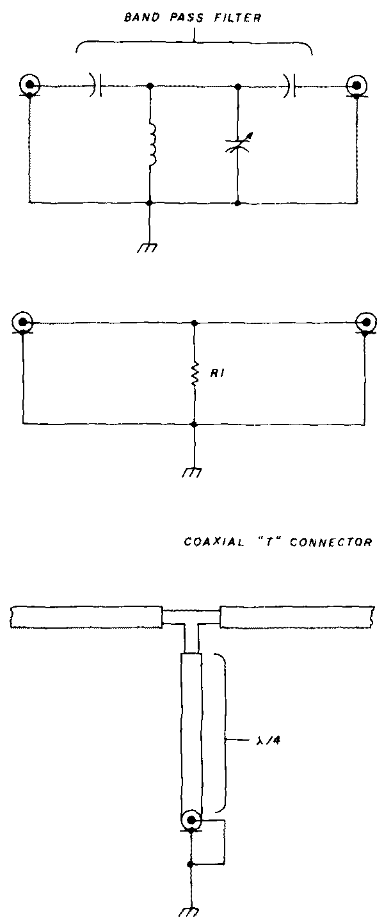

fig. 9. Other preamplifier protection schemes. R1 is typically 10 kilohm resistor.

components

Component abuse or overstressing is one of the biggest problems faced by Radio Amateurs. Many of the failures I hear about are not induced by lightning or by RF, but rather by the user who is trying to milk the last bit of performance from a device. *Solid state devices, especially those of the low-noise type, should never be operated at voltages, currents, or power dissipation levels beyond the manufacturer's specifications*! The importance of using protective diodes and zener diodes as well as current limiting resistors, as discussed above, cannot be overstressed. Likewise, all resistors should be sufficiently derated in power.

Capacitors, especially on a receiver input, should have adequate breakdown voltages. I've noticed that the CATV industry uses 1000 volts or higher breakdown voltages on all their capacitors on the input and output circuits. Surely they know something about lightning — they have thousands of miles of transmission lines all over the world!

summary

In this month's column, I have described many of the types of failure mechanisms that can plague an active VHF/UHF Amateur. They range from carelessness and unwise short-cuts to events beyond our control. This column is not meant to alarm you; in fact, the opposite is true. If you understand the limitations of your equipment and the outside stresses and forces that can be placed upon them, you can take adequate protective measures using some or all of the techniques mentioned in this column. Taking short-cuts on grounding, filtering, and switching is "penny-wise and pound foolish."

One final remark: some of the protection devices described may be destroyed if you're unfortunate enough to suffer a failure or lightning strike.

However, the cost of replacing these devices is usually far less than the cost of replacing the equipment they protect! Remember, *it's virtually impossible to protect yourself from a direct lightning hit but it is possible to minimize the destruction.*

acknowledgements

I'd like to thank Leroy May, W5HN, a great old-time VHF/UHFer, who, by his letter to me several years ago, encouraged me to write on this subject.

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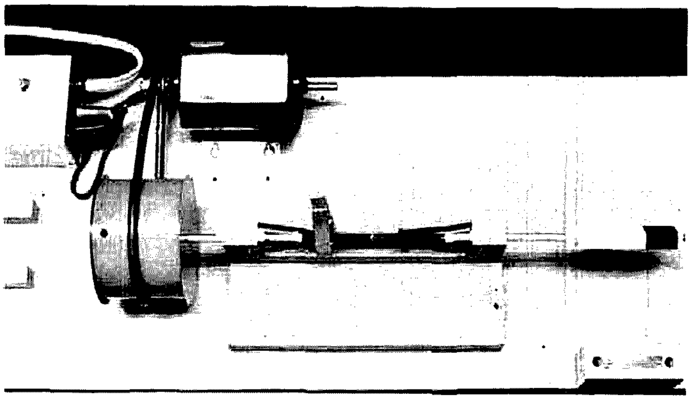
important VHF/UHF events:

June 1:	*EME Perigee*
June 5:	*Predicted peak of the Arietids meteor shower (1930 UTC)*
June 8-9:	*ARRL VHF QSO Party*
June 15:	*Predicted peak of the June Lyrids meteor shower (0400 UTC)*
June 21:	*Mean date for the two month annual peak of sporadic-E propagation*
June 29:	*EME Perigee*
June 29-30:	*SMIRK 6-meter Contest*
July 20:	*± two weeks, look for 2-meter sporadic E openings*
July 20-21:	*CQ VHF WPX Contest*
July 25:	*EME Perigee*
July 27-29:	*Central States VHF Conference, Tulsa, Oklahoma (Contact WØRRY/5)*
July 28:	*Predicted peak of Delta Aquarids meteor shower (0300 UTC)*

ham radio



Overall view of the coil winder set up to wind No. 30 enameled wire on a form approximately 0.8 inch (20 mm) × 0.43 inch (11 mm) × 2 inches (51 mm) long. About half of the winding length has been filled and the clothespin is clamped on the screw thread that causes the main shaft to traverse from left to right; the winding progresses from right to left. The mini-box at upper left is an SCR motor controller, and the two wood blocks at left are for later addition of a turns counter as explained in the text.

wind your own transformers — inexpensively

Novel approach produces high quality, low-cost replacements

I've seen many excellent articles on rewinding transformers in the home workshop. Valuable information for calculating the number of turns per volt relating to core size, insulation, varnish dipping, and space available was included, but in each of these articles it always seemed that in order to complete the project, it was necessary to somehow hold the coil in one hand and the wire in the other and just wind away. Many authors noted, of course, that the methods were really not suited to coils with a large number of turns of small wire, but were ideal for the larger sizes of wire, such as might be found powering the lower voltage transistor circuits popular in recent years.

This article describes a home-brew coil winding machine that will wind wire uniformly and neatly. It can be built from readily available, inexpensive parts. Most of the parts can be found right at home; the rest are available at reasonable prices. The only equipment necessary is a small drill press, a table saw, and the usual hand tools found around the hamshack. A 3 to 4 inch-circle cutter attachment for the drill press will save a lot of time.

background

Several years ago, when I experimented briefly with Amateur Television, I acquired a 12-year-old television camera with a bad power transformer. At first I thought it would be a simple matter to order a replacement transformer, but my enthusiasm faded when the manufacturer told me they'd stopped stocking parts years ago. My interest diminished even more when I learned that a professionally rebuilt replacement that would fit in the available space and meet the electrical requirements would cost $200.00.

My conscience wouldn't allow me to junk a perfectly good TV camera, and there was no way I'd spend more for a transformer than I'd invested in the whole camera. For years I'd toyed with the idea of a home-brewed coil winder; it seemed its time had finally come.

how it works

Small transformers are wound on a fairly complicated piece of machinery that rotates the form, or

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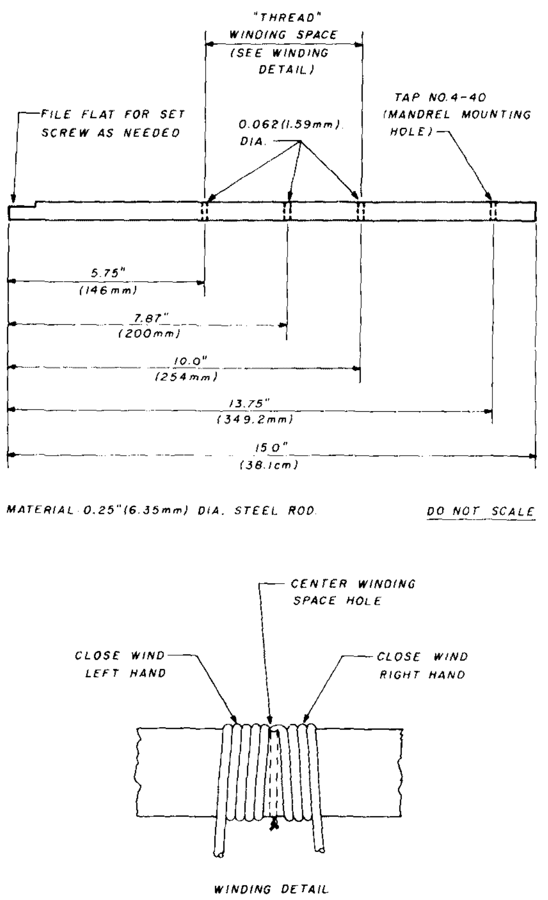


fig. 1. Dimensions of main shaft of coil winder.

mandrel, on which the wire is being wound. While the form makes one revolution, the feed guide moves the same distance as the diameter of the wire (plus perhaps a fraction of that distance for clearance). Such a winding lathe has a large selection of gears available so that the lead screw is turned at just the right rate to accomplish this motion. A transformer winding machine costs thousands of dollars — and the average ham obviously doesn't wind coils often enough to justify that expense.

It would be fairly easy for an Amateur to assemble a device that would merely rotate the coil form, but the mechanism necessary to move the wire guide at precisely the proper rate — unless you already have a lathe with many gears — gets much too complicated for a homebrew project.

In searching for ways to solve the problem inexpensively, it occurred to me that the shaft on which the coil form was mounted could do the lengthwise moving, while the wire feed point remained stationary. As I began to think about having lead screws made for each size of wire and the many other details, I realized I could use the wire itself for the lead screw! Normally you'd have a little bit of the size wire you were going to wind left over. So why not form the lead screw threads right on the shaft by simply winding on a single layer of wire for each direction of feed, securing it in place if necessary? It's fairly easy to close wind on a round rod, even with small wire; besides, you do it only once for each size of wire you're going to wind.

The shaft is made to move lengthwise while rotating by mounting a pair of pads of resilient material on strips of metal fastened to the bearing posts, close to the shaft but not touching, on either side. When feeding is desired, a clothespin is used to clip over the pads to bring them in contact with the "threads." There are two sets of these pads, one for each direction of feed, so when ready to reverse wire feed, you simply stop the rotation and move the clothespin to the other pair of pressure pads, insert and wrap the paper insulation layer over the layer just wound (securing it temporarily with masking tape if needed), and then restart winding with feed in the other direction. I designed my winder around 0.25-inch (6.35 mm) steel rod, and to accommodate an actual winding length of 1.8 inch (46 mm), which means that a 2-inch (51 mm) long winding form can be handled with a little space at each end. I wanted to have a winding shaft for each wire size to be used to save for use again with that particular size. This means, of course, that when you've finished winding one wire size, you remove the drive pulley and the winding form, mount them on another rod made up for the next wire size, and reinstall the unit into the bearings. The "bearings" are simply wood posts with a slot sawed into the top in which the winding shaft lies, leaving it free to slide lengthwise as well as to rotate.

The shaft is powered by an old sewing machine motor (complete with pulley, if possible) made to be variable in speed by the inclusion of an SCR speed controller built from a diagram in the General Electric Company SCR manual.[1] Similar circuits can be found in the handbooks: a plain old Variac — or even the foot control from the sewing motor — would do.

construction

Main shaft. The main winding shaft is made from 0.25-inch (6.35 mm) steel rod, with a flat filed on one end for the drive pulley set screw, three small holes for securing the "thread" winding, and a tapped hole for attaching the coil form and mandrel. (When buying the rod, be sure to select the straightest and smoothest you can find; its dimensions are shown in **fig. 1**.) It seems wise to postpone placing the winding on the shaft until the winder is complete enough to be used

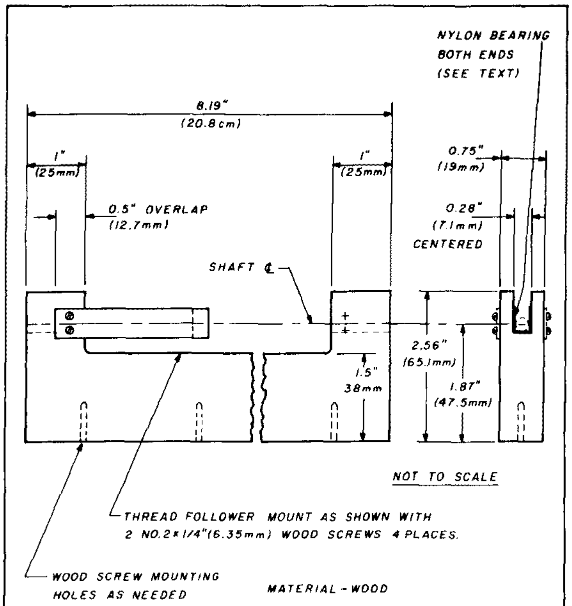


fig. 2. Bearing block details.

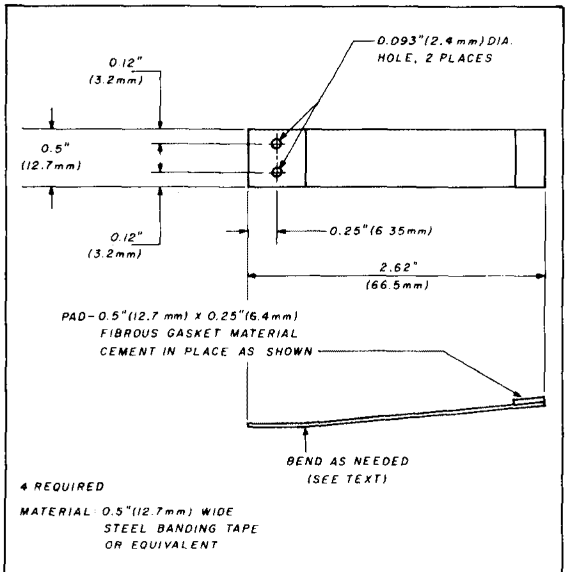


fig. 3. Thread follower details.

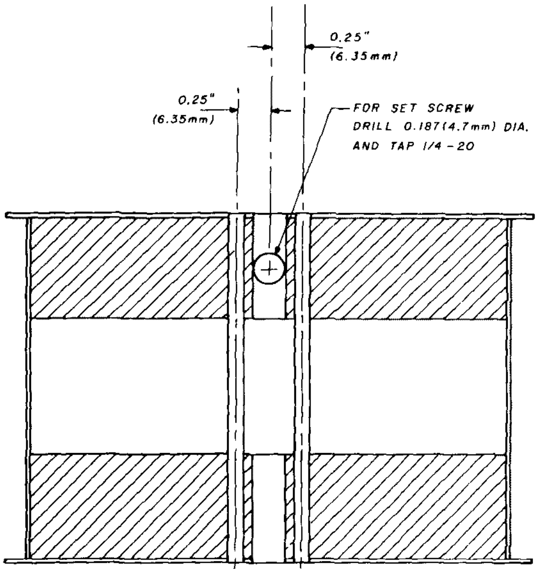
so that the motor can rotate the shaft for you. For the finer sizes — perhaps No. 30 and smaller — the wire can be cemented to the shaft to avoid an accident during winding. After cleaning with alcohol, I used a very thin coat of epoxy cement (not the "quick setting" type) applied to the winding space before winding on the wire. After winding, I carefully removed the excess from the outside of the winding with a cotton ball and denatured alcohol. The outside surface of the wire should be clean enough so that the follower pads can "feel" the individual turns.

Bearing and thread follower support. Dimensions for this part are shown in **fig. 2**. The bearings were constructed from a piece of ordinary 3/4-inch (1.90 cm) thick board. Note that the slots that form the bearings are slightly larger than needed to accept the main shaft rod; this is to permit lining the slot with several thicknesses of nylon fabric as needed to act as a low friction bearing. The shaft should not bind or be loose and wobbly, but should turn freely as well as slide lengthwise freely.

Screw thread follower. This part is made from ordinary 1/2-inch (13 mm) wide steel banding, such as used for strapping crates. If steel strapping isn't available, pieces can be cut from a tin can. Four pieces are mounted on the bearing posts so that they contact the wire "screw thread" when the clothespin is clipped over them. These are detailed in **fig. 3**, and the photo of the entire winder shows them clearly. The pads cemented to the pieces are important; they should be made from a resilient material that will "sink into" the grooves between the "screw thread" wires when pressed against them. I used automobile gasket material — the kind used for mounting water pumps and such. The note "bend as needed" in **fig. 3** means that after installing, a bit of judicious bending is in order so that no contact is made until the clothespin is clamped on. When it is, it should contact as many turns of the "thread" as possible.

Drive pulley. I fabricated this part by cutting two discs from 5/8 inch (15.9 mm) particle board using the circle cutter in the drill press, truing them up by mounting them on a 0.25-inch bolt in the drill press, as if it were a lathe. I turned the cutting bit in the circle cutter so the disc being cut had straight sides, not the hole. (In the absence of a circle cutter, a sabre saw would probably do the job if, after cutting, you then mount the discs in the drill press and straighten up their sides with a sanding block — they should run as true as possible. I slipped the two pieces on a rod (for alignment), spaced for the width of the drive pulley, and, using epoxy, then cemented the thin aluminum around them to form a drum, using several rubber bands to hold them until the epoxy set. Then the flanges were cemented onto the ends of the pulley. (Incidentally, if a circle cutter is used in the drill press, clamp the work piece down securely and be sure to follow all the safety rules.)

Both the drum surface and the end flanges were cut from discarded aluminum offset printing plates. Although only between 0.010 inch (0.25 mm) and 0.012 inch (0.30 mm) thick, this metal works nicely. It should be available inexpensively from print shops or news-



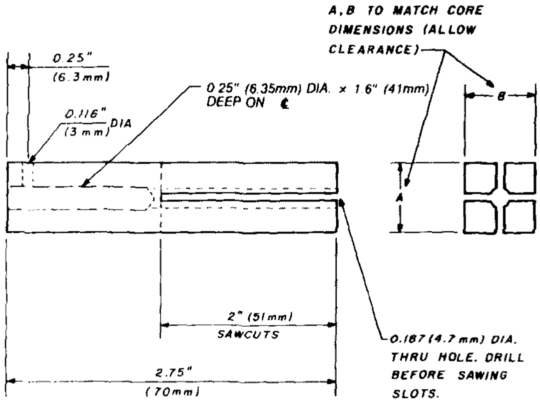
2.5 × 11.5 inch (6.35 × 29.2 cm) of 0.010 to 0.012 inch (0.25 to 0.3 mm) thick aluminum sheet (used printing plate). Position end disks on 0.25 inch (6.3 mm) rod for alignment. Wrap around end disks, cemented with epoxy (including overlap). Secure with sufficient rubber bands until set. Align carefully so aluminum sheet does not overhang disks.

End disks (2 needed): Make from particle board or plywood, 0.6 to 0.75 inch (15 to 19 mm) thick. Diameter: 3.5 inch (89 mm) with 0.25 inch (6.35 mm) hole in center.

2 holes all the way through, 0.125 inch (3.2 mm) diameter. Insert 2.5 inch (6.35 mm) long × 0.125 inch (3.2 mm) OD brass tubing and cement (for later addition of a turns counter). Drill these after assembly is complete.

2 flanges needed, 3.87-inch (98 mm) diameter with 0.25 inch (6.35 mm) hole in center. Make from aluminum sheet or thin, stiff cardboard. Cement to end disks.

fig. 4. Dimensions for drum parts and fabrication.



Select stiff, straight wires at least 5.5 inches (14 cm) long and small enough to slide freely in the brass tubing lined holes through the drive pulley (old bike spokes).

Position counter so that ends of counter drive wires are about 0.125 inch (3.2 mm) from nearest side of bearing block.

Build up counter shaft with tape. Place wires on opposite side and tape wires in place so their centers are 0.5 inch (12.7 mm) apart.

Make counter support block so that center of counter shaft is at same height as center of winding shaft and aligned with it. Block should be removable to enable changing rod.

fig. 5. Coil Mandrel.

papers. Since the drive belt should never reach the edge of the pulley, the flanges aren't really necessary, but they look nice.

The hole for the set screw that secures the pulley on the shaft is tapped edgewise in the particle board (1/4-20), and so far the threads haven't stripped out. As an additional feature, I installed two pieces of 1/8 inch (3.27 mm) OD brass tubing (cut from discarded ball point pen refills) opposite each other on about 0.5 inch (12.7 mm) diameter and lengthwise through the pulley (avoid the set screw!). This was to provide the addition of a counter with a couple of stiff wires taped to its shaft that could be inserted inside these holes and slide freely as the shaft moved back and forth. Dimensions for the drum parts and fabrication are shown in **fig. 4**. If one should want to add a counter as mentioned above, a sketch showing how I did it is shown in **fig. 5**.

Base and layout. I used a piece of 1/2 inch (12.7 mm) plywood, thickened to 1 inch (25.4 mm) around the edges, measuring approximately 26 × 10 inches (66 × 25 cm), as a base for the entire coil winder. Any convenient layout is satisfactory. I mounted the shaft bearing against an additional piece of wood the same length, using small nails and glue, providing a "lip" that could be mounted from the top side. Otherwise, mounting with wood screws up from the bottom side of the base into the bottom edge of the bearing should do the job. The drive belt won't be specified, but many are available that will work. The motor should be positioned so that the belt is just tight enough to run without slipping. Too tight a belt will cause obvious problems. A sewing machine shop would be the best place to look for a suitable drive belt; other sources might include jewelers or other craftsmen who could supply the belting used to run smaller lathes. This type of belting comes in a continuous length. A suitable piece is cut and the ends are joined together by heating.

The direction of rotation should be considered. In

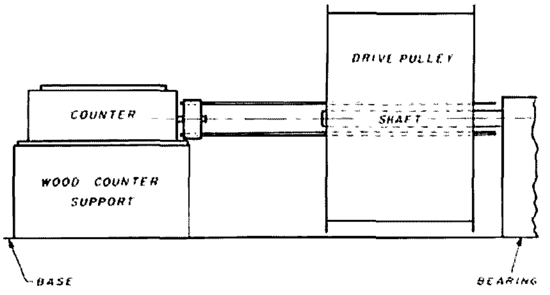


fig. 6. Method of coupling counter to winding shaft while still retaining freedom of lateral shaft movement.

my opinion the top of the shaft should be moving away from you when you stand in front of the winder; this facilitates inserting insulation papers and observing the wire as it winds on.

Wire tension and feeding. In keeping with the idea of building something for practically nothing, the winding wire is fed onto the coil over a block of wood (see lower right in photo of completed unit) between two pieces of felt clamped in place by a short strip of metal held in position by small wooden screws. Glue one piece of felt to the piece of metal and one to the top of the wood block, and adjust wire tension by tightening or loosening the screws. Tension should be maintained as uniformly as possible all during the winding of a coil; too much tends to "crunch up" the layers already wound underneath, and too little leaves the wire already wound free to move around or slip out of position. A much more elaborate wire tensioning system would be preferred, but this method will hold the wire in position accurately enough for now. Feeding the wire off the end of the supply spool helps to avoid any additional drag.

Winding mandrel or form. I made this of wood cut to match the core size of the transformer, and with a hole diameter in one end to fit on the winding shaft, and in the other to accept a No. 10 screw (push in), and with saw cuts so that when the screw is removed, the mandrel shrinks a little to allow coil removal. To be safe, I included some paper shims — when the coil is finished, it's too late to discover that it won't go over the core! (See **fig. 6** for details.)

General. Once this winder is in operation, it will become apparent that the lateral position of the shaft is rather delicate, so be careful to see that nothing gets in its way while winding. Once a set-up was made, I tested it by having the shaft run back and forth several times under power and with the follower pads clamped on just to make sure everything was in order before actually starting the winding. This also helps the follower pads to match the shape of the wire screw thread.

A thin coat of lubricant such as petroleum jelly on the screw thread winding is suggested.

One more precaution should be mentioned. Since not all enameled wire of a given copper size has the same thickness of enamel on it, the best course to follow is to use wire from the same spool for both the screw thread winding and the coil, if not, make sure the wire for the thread is *not smaller* in overall diameter than that used for the wire to be wound.

I was able to wind over 2300 turns of No. 42 enameled wire on a coil without any problems, and found that it could be run at over 200 RPM as long as the speed was brought up gradually. (In case you're wondering, I was able to rewind the power transformer for the TV camera — and it's still working!)

This design could easily be adapted to provide for greater winding lengths than the 1.8 inch (46 mm) I have provided for. Just remember to increase the space between the bearing posts and the overall length of the shaft rod to allow for the extra lateral movement needed.

references

1. *SCR Manual*, 4th Edition, General Electric Company, page 210.

ham radio

DX FORECASTER

Garth Stonehocker, K0RYW

sporadic-E propagation

During the summer months, the sun, almost directly overhead, produces more ions in the lower ionosphere than it does in winter. These ions support short-skip propagation — even *multiple* short-skips. The geomagnetic field clusters and forces these overly abundant ions into cloud-like patches known as sporadic-E (E_S). These patches, which form in a thin-dense layer about 60 miles (100 km) above the earth, give rise to strong, mirror-like signal reflections over short-skip distances of 600 to 1200 miles (1000 to 2000 km).

Because E_S is related to the summer sun, the best locations for working these E_S openings are in the Northern Hemisphere from June through September and in the Southern Hemisphere during its summer, December through March. In each hemisphere the best E_S is found on either side of the geomagnetic equator; it's especially good where the geomagnetic equator is farthest from the geographic equator, giving greater distance and force for formation. These special areas are South-east Asia in the Northern Hemisphere and South America in the Southern Hemisphere, with the former the better of the two. This is because the sun's maximum level of ion production occurs in the D and E layers (37 to 62 miles, or 60 to 100 km) above the earth at 23 degrees latitude, directly under the sun. As the ionization travels (in movement known as *diffusion* or *drift*) from this maximum to less dense areas, E_S is bunched by field-strength variations. When ions move, they can move only parallel to the geomagnetic field lines, not perpendicular to them. (See cross-section views of the earth's magnetic field lines in **fig. 1**, particularly those lines connecting the northern and southern hemispheres.) **Figure 2**[1] shows the movement of E_S patches across the United States. The E_S patches are the variation (bunching) of ions caused by changes in the geomagnetic field strength modulating and forcing the ions in their north-south movement. In these two special areas of maximum separation between equators, the modulating force and the distance over which to bunch are greatest, so more E_S patches are formed.

last-minute forecast

DX conditions for the higher frequency bands, 10 through 30 meters, are expected to be very good during the first and last weeks of the month when the solar flux could be high. (Verify this daily by checking WWV at 18 minutes after each hour.) During the middle weeks of the month, the lower frequency bands should engage interest for some short skip daytime and DX at night. Geomagnetic disturbances are expected on June 6th and 16th. The moon will be full on the 3rd, and closest approach (perigee) will

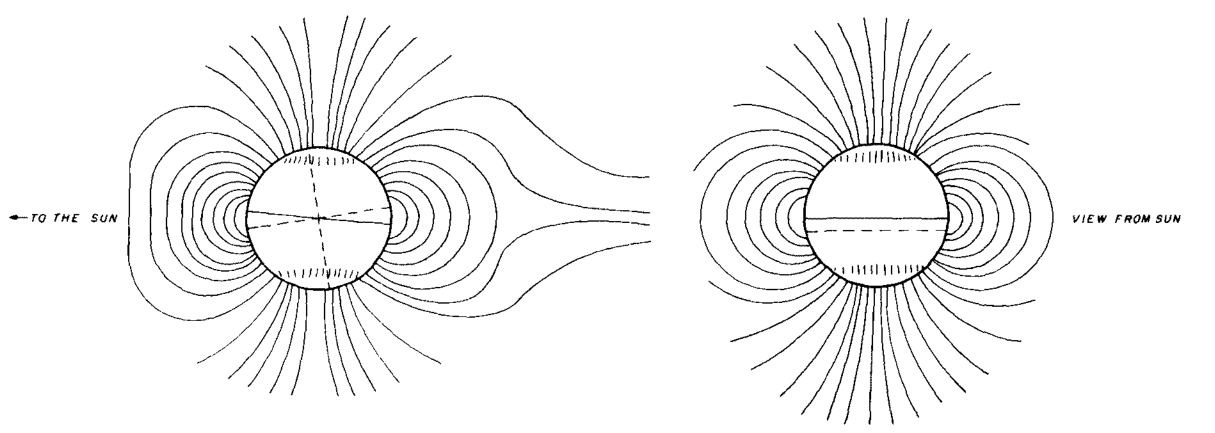


fig. 1. The earth's magnetic dipole field, depicted relative to the size of the earth, shows that ions can move only north or south along field lines, not east or west across them. Consider, for example, the movement of E_S patches across the U.S. as shown in figure 2.[1]

JUNE

WESTERN USA

GMT	PDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	5:00	20	20	15	15	20	10	15	20
0100	6:00	20	20	20	15	20	10	15	20
0200	7:00	20	20	20	15	20	10	15	20
0300	8:00	20	20	20	15	20	15	15	20
0400	9:00	20	20	20	15	30	15	15	20
0500	10:00	20	20	20	15	30	15	15	20
0600	11:00	20	30	20	15	30	15	15	20
0700	12:00	20	30	20	15	*30*	20	15	20
0800	1:00	20	*30*	30*	20	*30*	20	20	20
0900	2:00	20	*30*	*20*	20	*30*	20	*20*	20
1000	3:00	20	*30*	*20*	*20*	*30*	*20*	*20*	20
1100	4:00	*20*	20	*20*	*20*	*30*	*20*	*20*	*20*
1200	5:00	*20*	20	15	*20*	*30*	*20*	*20*	*20*
1300	6:00	*20*	20	15	*20*	*30*	*20*	*20*	*30*
1400	7:00	*20*	20	15	*20*	*30*	*20*	*20*	*30*
1500	8:00	*20*	20	15	_30*_	30	*20*	*30*	*30*
1600	9:00	*20*	20	15	*30*	30	20	20	*30*
1700	10:00	*20*	20	15	20	20	15	20	20
1800	11:00	*20*	20	15	20	20	15	20	20
1900	12:00	*20*	20	10	20	20	15	15	20
2000	1:00	20	20	10	15	20	15	15	20
2100	2:00	20	20	15	15	20	15	15	20
2200	3:00	20	20	15	15	20	15	15	20
2300	4:00	20	20	15	15	20	10	15	20

MID USA

GMT	MDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)	CDT
0000	6:00	20	20	15	20	20	10	15	20	7:00
0100	7:00	20	20	20	20	20	10	15	20	8:00
0200	8:00	20	20	20	20	30	10	15	20	9:00
0300	9:00	*30*	20	20	20	30	15	15	20	10:00
0400	10:00	*30*	20	20	20	30	15	15	20	11:00
0500	11:00	*30*	20	20	20	30	15	15	20	12:00
0600	12:00	*30*	30	20	20	30	20	15	20	1:00
0700	1:00	*30*	*30*	20	20	*30*	20	20	20	2:00
0800	2:00	*20*	*30*	20	*30*	*30*	20	*20*	20	3:00
0900	3:00	*20*	*30*	*20*	*20*	*30*	20	*20*	20	4:00
1000	4:00	*20*	*30*	*20*	20	*30*	*20*	*20*	20	5:00
1100	5:00	20	20	*20*	20	*30*	*20*	*20*	*20*	6:00
1200	6:00	20	20	15	15	*30*	*20*	*20*	*20*	7:00
1300	7:00	20	20	15	15	*30*	*20*	*20*	*20*	8:00
1400	8:00	20	20	15	15	*30*	*20*	*20*	*20*	9:00
1500	9:00	20	20	15	15	30	*20*	*20*	*20*	10:00
1600	10:00	20	20	15	15	30	20	20	*20*	11:00
1700	11:00	20	20	15	15	30	15	20	20	12:00
1800	12:00	20	20	10	15	20	15	20	20	1:00
1900	1:00	20	20	10	15	20	15	20	20	2:00
2000	2:00	20	20	10	15	20	15	15	20	3:00
2100	3:00	20	20	15	15	20	15	15	20	4:00
2200	4:00	20	20	15	15	20	15	15	20	5:00
2300	5:00	20	20	15	15	20	10	15	20	6:00

EASTERN USA

GMT	EDT	N (Asia Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania Australia)	NW (Japan)
0000	8:00	20	20	15	20	20	10	15	20
0100	9:00	20	20	20	20	20	10	15	20
0200	10:00	20	20	20	20	30	10	15	20
0300	11:00	20	20	20	20	30	15	15	20
0400	12:00	*30*	20	20	20	30	15	15	20
0500	1:00	*30*	20	20	20	30	15	15	20
0600	2:00	*30*	20	20	20	30	20	15	20
0700	3:00	*30*	*20*	20	30*	*30*	20	20	20
0800	4:00	*20*	*20*	30*	*30*	*30*	20	*20*	20
0900	5:00	*20*	*20*	*20*	*20*	*30*	20	*20*	20
1000	6:00	*20*	*20*	*20*	20	*30*	*20*	*20*	20
1100	7:00	20	20	20	20	*30*	*20*	*20*	*20*
1200	8:00	20	20	15	15	*30*	*20*	*20*	*20*
1300	9:00	20	20	15	15	*30*	*20*	*20*	*20*
1400	10:00	20	20	15	15	30	*20*	*20*	*20*
1500	11:00	20	20	15	15	30	*20*	*30*	*20*
1600	12:00	20	20	15	15	30	20	20	20
1700	1:00	20	20	15	15	30	15	20	20
1800	2:00	20	20	10	15	20	15	20	20
1900	3:00	20	20	10	15	20	15	20	20
2000	4:00	20	20	10	15	20	15	15	20
2100	5:00	20	20	15	15	20	15	15	20
2200	6:00	20	20	15	15	20	15	15	20
2300	7:00	20	20	15	15	20	10	15	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

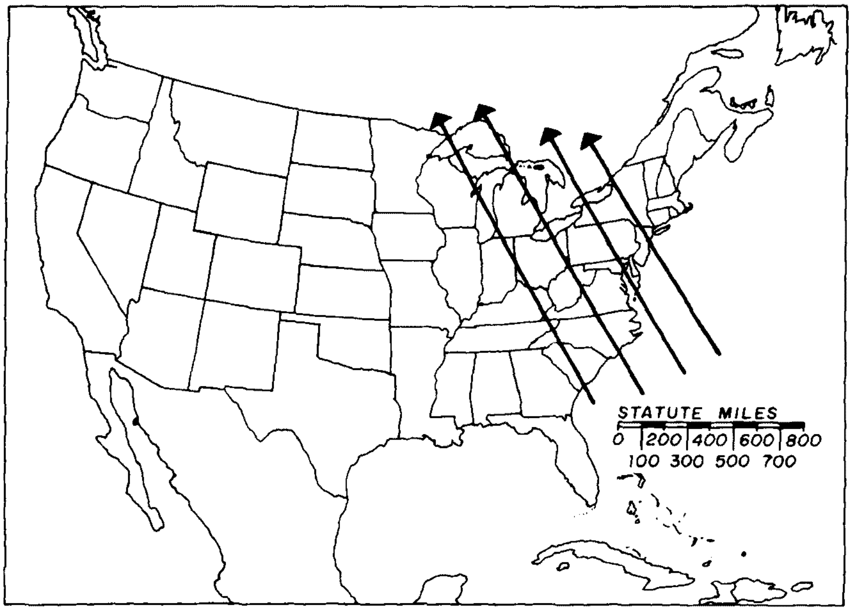


fig. 2. Clouds producing sporadic-E propagation generally travel from southeast to northwest at approximately 180 miles (280 km) per hour, moving in a relatively straight line.

occur on the 1st and 29th. Summer solstice is on the 21st at 1044 UTC. The Aquarid meteor shower starts about the 18th, peaks about the 28th, and lasts until about August 7. The maximum radio-echo rate will be 34 per hour.

band-by-band summary

Six meters will provide occasional openings to South Africa and South America around noontime by short-skip E_s.

Ten meters will provide long-skip conditions in the afternoon during the peak times of the 27-day solar cycle. Otherwise, look to sporadic-E short-skip and multihop openings around local noon for DX on these bands. (Transequatorial evening openings do not usually occur in the summertime.)

Fifteen and twenty meters, almost always open to some part of the world, will be the main daytime DX bands. Twenty meters will stay open on long southern paths into the night, though 15 will drop out in the late afternoon. Operate on 15 first, then move down to 20 meters later. DX is 5000 to 7000 miles (8000 to 11,200 km) on these bands. There may be some long one-hop transequatorial propagation.

Thirty and forty meters are both daytime and nighttime bands. Intermediate distance operation (1000 to 1500 miles, 1600 to 2400 km), in any direction is considered daytime DX. Nighttime DX on these two bands may be expected to occur over greater distances than on 80 meters and, like 80, will follow the darkness path across the sky. Signal strength and distances covered are reduced on days of high solar flux values. In addition, no 30-meter openings will take place during the pre-dawn hours on the morning after these high radio flux values.

Eighty and one-sixty meters will exhibit short-skip conditions during daylight hours and lengthen for DX near dark. Eighty meters will open to the east just before your sunset swing more to the south as midnight approaches, and end up in the Pacific areas during the hour or so before dawn. (One-sixty opens later and ends earlier.)

references

1. Joe Reisert, W1JR, "The VHF/UHF Primer: An Introduction to Propagation," *ham radio*, July, 1984, page 14.

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the ham notebook

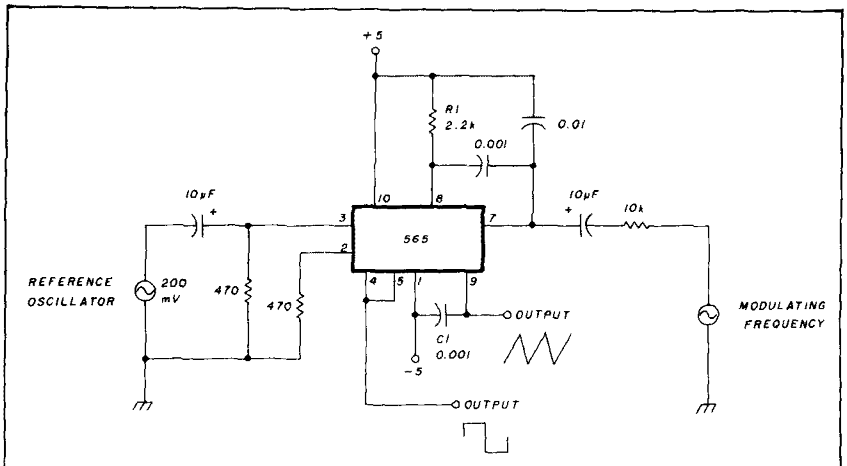

fig. 1. Schematic diagram of the experimental PLL phase modulator.

PLL phase modulator

I devised this circuit to help students understand the difference between frequency and phase modulation. It is also useful for studying the effects of phase-shift behavior in a closed-loop circuit.

Most phase modulators contain a certain amount of amplitude modulation, but this one does not. Its basic circuit, shown in **fig. 1**, uses a 565 PLL chip as a modulator, which is limited to a maximum frequency of about 500 kHz.

Resistor R1 and capacitor C1 establish the approximate operating modulation frequency where $f \cong 1.2/4R1C1$. The reference or input frequency should be adjusted until lock is obtained as described below. If desired, the input signal can be provided by a crystal-controlled oscillator (this is one of the chief advantages of phase modulation).

Lock can be determined by observing the input frequency on one channel of a dual-trace scope and the VCO frequency on the other channel. Pin 9 of the 565 should be used to observe the VCO frequency because a triangular waveform is available at this point. Pins 4 and 5 provide the normal square-wave outputs of the chip, but the fast rise and fall times make it hard to see what is happening to the waveforms at this level.

Once lock is achieved, the input frequency should be adjusted until the reference oscillator frequency and the VCO frequency are in phase. Alternatively, the value of R1 and C1 can be adjusted to shift the phase of the VCO signal.

The next step is to apply the modulating frequency to the error voltage input pin. A very low frequency should be used at first. Observe the display on an oscilloscope, making sure the scope is triggered on the reference signal. If the circuit is working correctly, the reference signal should be steady and the VCO signal should shift slowly back and forth in phase. The VCO is now being phase modulated.

Try experimenting with different amplitudes of the modulating frequency. The amount of modulation should vary with the amplitude of the modulating frequency. If the modulation amplitude is too high, the VCO may lose lock.

The amount of modulation is somewhat limited in this circuit. If more modulation is needed, additional modulators may be added in cascade. The ouput of the first modulator must be taken from pin 4 or 5 and reduced in level to about 200 mW before being fed to the second modulator.

Another characteristic of phase modulators may also be observed. As the modulation frequency is increased, the amount of modulation also increases. (This is another advantage of a phase modulator over an FM modulator.) Pre-emphasis is built right into the circuit and does not have to be added externally.

I have found this circuit easy to build and helpful to students. It can be converted to FM by removing the reference oscillator.

I would be interested in hearing from other *ham radio* readers about applications using other chips at higher frequencies.

bibliography

Berlin, Howard M., *The Phase-Lock Loop Bugbook, with Experiments*, page 7-7 to 7-10. Capital City Press, Montpelier, Vermont, 1978.

Graham W. Stratford, VE3FHM

short circuits

June 1985

MINIMUF modification

The software code modification of the MINIMUF 3 propagation program provided in **fig. 2** of the February, 1985 "DX Forecaster" (page 75) contains an error. **Line 1774**, which reads **IF F2>F3 THEN 1779** should read **IF F2>F3 THEN 1780.**

June 1985

filter design

The program shown in N6JH's "Computer-aided Interdigital Bandpass Filter Design" (January, 1985, page 12) will correctly execute all examples given in the text and are applicable to any Chebyshev filter design. However, if a Butterworth design is desired (0dB entered as the desired ripple in the passband), the program will plot the attenuation graph incorrectly, although it will compute the mechanical details correctly. Changing **line 970** from **GOTO 1040** to **GOTO 1020** will correct this problem.

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WANTED: National NC-101X. Will buy for cash or trade nice HRO Sr. For sale 75A4 Ser. No. 1479 $225. Prefer pick-up. Horace Goss, W2AB, Crosstrees Hill, Essex, CT 06426. Tel 203-767-8485.

WANTED: Mono and early stereo tube HiFi equipment. Marantz, Scott, Fisher, Eico, McIntosh, Dynaco, etc. Working or not, complete or parts. Jack Smith, 59 Mill Pond, N. Andover, MA 01845.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

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RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

RETAIL STORE being liquidated by mail auction. Parts, tools, test equipment, home supplies. SASE to ESI, Box 328, Mesa, AZ 85201.

Coming Events

ACTIVITIES

"Places to go..."

ILLINOIS: The DuPage Amateur Radio Club Hamfest/Computerfest, Sunday, July 14, Downers Grove American Legion Post 80. Gates open 8 AM. Tickets $3.00 available at the gate only. Large outdoor flea market and swappers row. Inside exhibits available. VEC exams. Plenty of parking. Food and drink available. Talk in on 146.52 simplex. For information SASE to W9DUP, PO Box 71, Clarendon Hills, IL 60514 or call (312) 971-3294 8 AM to 9 PM.

PENNSYLVANIA: Harrisburg Annual Firecracker Hamfest, Thursday, July 4. Sponsored by the Harrisburg RAC, Bressler Fire Co. picnic grounds, Exit 1 of I 283. Nearby motels and restaurants. Plenty of parking. Shaded pavillion with tables. Free tailgating. Admission $3.00. XYL and kids free. A test session for hams wishing to upgrade will be held nearby. Send check for $4.00 payable to ARRL-VEC with FCC form 610. Some limited walk-ins. For more details and table reservations: Dave, KC3MG, 131 Livingston St, Swatara, PA 17113 or (717) 939-4957.

OHIO: The 21st annual Wood County Ham-A-Rama, Sunday, July 14, Wood County Fairgrounds, Bowling Green. Gates open 8 AM. Free admission and parking. Trunk sales. Food available. Advance table rentals $5.00 to dealers only. Saturday setup until 8 PM. K8TIH talk in on 147.18 repeater and .52. For more information or dealer rentals SASE to Wood County ARC, c/o Craig Henderson, N8DJB, 7368 Scotch Ridge Rd., Pemberville, OH 43450.

MICHIGAN: The "85" U.P. Hamfest, July 27 and 28, St. Francis de Sales School, Manistique. Friday evening Fish Fry, set and eyeball for early arrivals. Saturday 6 AM to 5 PM. Banquet 6:30 PM. Sunday 8 AM to 2 PM. Registration $3.50. Free baby sitting. Table space $3.00 per 4' table. For more information: Debbie Barton, WD8IBT, 509 Range St., Manistique, MI 49854. (906) 341-5694 after 3 PM.

PENNSYLVANIA: The MURGAS ARC (K3YTL) will sponsor the annual Wilkes-Barre Hamfest, Sunday, July 7, rain or shine, 109th FA Armory, Market Street, Kingston (across the river from Wilkes-Barre). Set up 6 AM. General admission 8 AM. Registration $3.00. Women and children under 16 free. Tailgating $2.00 per space. Tables and commercial power available. Talk in on 146.01/.61 and 52 simplex. For further information write: Hamfest Committee, PO Box 1094, Wilkes-Barre, PA 18703 or call (717) 388-6863.

BRITISH COLUMBIA: The 33rd annual Pacific Northwest DX Convention, sponsored by the British Columbia DX Club, Saturday and Sunday, July 27 and 28, Richmond Inn, Richmond, British Columbia. DXers from all over the world will be here to visit, learn and swap stories about our favorite hobby. Send club rosters or labels for mailing to individual club members for further details about the convention. The labels will be used only for this purpose. See you in British Columbia! Ken Thompson, VE7BXG, Chairman, 12467 - 53rd Avenue, Surrey, BC V3W 1A4 or Andy Ponzini, VE7AHA, Publicity, 4551 Arthur Drive, Delta, BC V4K 2X4.

INDIANA: State ARRL Convention and Indianapolis Hamfest, Saturday and Sunday, July 13 and 14. Marion County Fairgrounds, I 74 and I 465. Flea Market, commercial vendors, free camper facilities and hookup available on grounds. Motels nearby. Gates open 6 AM. Tech forums, ARRL convention and banquet. Food service on grounds. Gate ticket $5.00 gets free parking and more. For further information: Indianapolis Hamfest, PO Box 11776, Indianapolis, IN 46201.

1985 BLOSSOMLAND BLAST, Sunday, October 6, 1985. Write "BLAST", PO Box 175, St. Joseph, MI 49085.

MAINE: The YL International Sideband System's annual Convention, June 27-30, Sugarloaf/USA, Kingfield. Accommodations are available for reasonable rates. RV parking. Besides the regular meetings, DX forum, etc there are planned tours of Rangeley Lake area and Sugerloaf/USA. For complete details and registration packet send business SASE, 37¢ to: Phyllis Davis, KA1JC, PO Box 805, Presque Isle, ME 04769.

NEW YORK: The Mt. Beacon Amateur Radio Club's Hamfest, Saturday, July 13, Arlington Senior High School, Poughkeepsie/Lagrange. Note date change. Doors open 8 AM. Tickets $2.00. XYL and children free. Tailgating space $3.00 (1 free admission). Table $4.00 (1 free table and admission). Auction 2 PM. Talk in on 146.37/97 and 146.52. For information: Julius Jones, W2IHY, RR2, Vanessa Ln, Staatsburg, NY 12580 or call (914) 889-4933.

INDIANA: The combined LaPorte-Michigan City Amateur Radio Clubs will sponsor Summer Hamfest, Sunday, July 14, at the LaPorte County Fairgrounds, State Road 2, West of LaPorte. 8 AM to 2 PM. Donation $3.00 at the gate. Paved parking. Indoor tables by reservation 40¢/ft to Box 30, LaPorte, IN 46350.

WISCONSIN: The Aug Claire Amateur Radio Club will hold its annual Hamfest, Saturday, July 13, 4-H Buildings, Eau Claire. 8 AM to 4 PM. Free tables and coffee. Tickets $2.00 in advance, $3.00 at door. Talk in on 31/91 and 52 simplex. For info, tickets, SASE to Gene Lieberg, KA9DWH, 2840 Saturn Avenue, Eau Claire, WI 54703.

OHIO: The Lancaster and Fairfield County Amateur Radio Club's annual Hamfest, Sunday, July 14, Fairfield County Fairgrounds. 8 AM to 4 PM. In memory of WB8VOA. Admission $3.00 advance, $4.00 at door. Tables $4.00 advance, $5.00 at door. Table space $3.00 advance, $4.00/door. Talk in on 147.03/63 or 146.52 simplex. Refreshments available. Plenty of parking. For more information write Lancaster ARC, Box #3, Lancaster, Ohio 43130.

OHIO: The 11th annual Hall of Fame Hamfest presented by W8ZX, Tusco ARC and W8AL, Canton ARC, July 14, Nimishillen Grange, 6461 Easton Street in Louisville. Registration $2.50/adv, $3.00/door. Mobile check in 146.52/52 and 147.71/12 call W8ZX. Large flea market, dealers, forums and more. Flea market parking $2.00 per vehicle. Deadline for table reservations is July 1. For more information or reserva-

tions: WA8SHP, Butch Lebold, 10877 Hazelview Ave., Alliance, Ohio 44601. (216) 821-6794.

WISCONSIN: The Oshkosh Amateur Radio Club in conjunction with the S.O.L.A.R. Assn. will host EAA hams for the 1985 convention, July 26 - August 2. Stop and rest, charge your batteries, leave messages, etc at the EAA Ham Shack located at the north end of the commercial exhibit area. Look for the red and white ARRL flag. On Saturday, July 27 at 3 PM, there will be a gathering for all EAA hams hosted by the Oshkosh ARC. We'll be serving bratwurst, burgers and refreshments free of charge. Bring your wives and kids. You're in for a treat! For further info: Forest Schafer, WD9IWL, 417 Willow St., Omro, WI 54963.

KENTUCKY: The Northern Kentucky Amateur Radio Club's "Hamorama '85", June 15 and 16, Best Western Vegas Convention Center, Erlanger. All indoors, air-conditioned and free parking. Vendor set up Friday evening, June 14 after 8 PM. Flea market set up after 6 AM both Saturday and Sunday. General admission 8 AM. Admission for both days $5.00 per person. Entire family for $8.00. Children under 16 free. Flea market tables $5.00 each for entire weekend. Talk in on 147.86/.26 or 147.975/.375. For information: John A. Themes, WM4T, VP, NKART, 60 Locust Avenue, Covington, KY 41017. (513) 397-7425 days or (606) 331-0331 evenings.

MONTANA: The EYARS of Eastern Montana will hold the annual Fathers Day Picnic, June 16, National Guard Armory at the Glendive Fairgrounds. Registration at 8 AM. Polluck at 1 PM. Licensing exams are tentative pending interest. Camping hookups available. For information contact Dave Bruenı, KC7AA, 215 Third St. H.P., Glendive, MT 59330.

WASHINGTON: The Apple City Amateur Radio Club's Hamfest, June 8 and 9, Rocky Reach Dam, 7 miles north of Wenatchee. Registration: Amateurs $5.00; others $1.00; under 12 free. Banquet dinner $7.00 per person by June 3. Talk in 146.07/67 or 146.49 simplex. For motel/dinner reservations contact any Wenatchee ham.

NEW JERSEY: The Raritan Valley Radio Club will hold its 14th annual Hamfest, Saturday, June 15, Columbia Park, Dunellen. Gates open 8:30 AM. Sellers spots $5.00 each, no tables supplied. Lookers $2.00 donation. Refreshment stand. Talk in on club repeater, W2QW/R 146.025/.625 and 146.52 simplex. Advance tickets may be purchased from any club member. For further information call Jack, W2IWK at (201) 756-2546 or Ted, WB2TKU (201) 725-3481 between 10 AM and 10 PM.

CALIFORNIA: The Satellite Amateur Radio Club will have its annual Swapfest and Barbeque, Union Oil Picnic Grounds, south of Santa Maria, on Father's Day, June 16. General admission 9 AM and the Barbeque will be served at 1 PM. Dinner and tickets adults $7.95, children 6-12 $3.50. Children under 6 free. Swap spaces $3.50 each. For information, tickets or reserved tables write: Satellite ARC Swapfest, PO Box 1753, Santa Maria, CA 93456.

NEW HAMPSHIRE: *Fly in to New Hampshire's 2nd* largest Amateur Radio/electronic flea market at Manchester Municipal Airport, Saturday, June 29. Starts 9 AM. Sponsored by the New Hampshire FM Association. Rain date Sunday, June 30. General admission $1.00 per person. Sellers $5.00 bring own table or tailgate. Commercial displays welcome. Refreshments available. Talk in on 146.52 FM. For further information Doug Aiken, K1WPM (603) 622-0831 or Peat Henriksen, WA1RCF, 123 Woodlawn Circle, Portsmouth, NH 03801 (603) 431-5432.

WYOMING: The Great Plains Repeater Association and the High Plains Amateur Radio Club are jointly sponsoring the 1985 *Wyoming Hamfest, July 12, 13, and 14*, Wyoming State Fairgrounds in Douglas. Distributor displays, indoor flea market, tables available. License exams, seminars, auction, banquet, breakfast and more. Ample RV parking w/out hookups. Plenty of motels. For information and money saving advance registration SASE to Doug DesEnfants, WA7WXO, North Star Route, Torrington, WY 82240.

OREGON: The Northwest Division of the Mercury Amateur Radio Association will host the first annual MARA convention at Camp Ester Applegate, near Klamath Falls, June 20, 21 and 22. MARA will operate special event station W7UFM in conjunction with this event from 2000Z, June 20 to 2400Z June 22. For a special commemorative QST send large SASE to MARA, c/o Jack Jakoubek, KD7EZ, 477 Deep Creek Road, Chelhalis, WA 98532.

OHIO: The Magic Valley Chapter of the Idaho Socity of Radio Amateurs will host a Swapmeet on Saturday, June 15 from 9 AM to 5 PM at the Moose Lodge, 835 Falls Avenue, Twin Falls. Free admission. Swap tables $2.00. All indoors. FCC exams and ARRL reps. Talk in on 16/76. For further info write to PO Box 294, Twin Falls, ID 83303.

VIRGINIA: The Old Virginia Hams ARC announces its 11th annual Manassas Hamfest, June 2, Prince William County Fairgrounds, off Rt 234 1/2 mile south of Manassas. Gates open 8 AM. Tailgate setup 7 AM. Admission $4, under 12 free. Tailgaters $3.00 per space extra. YL program. CW proficiency awards. For information: Art Whittum, W1CRO, c/o Ole Virginia Hams ARC, PO Box 1255, Manassas, VA 22110. Tel (703) 361-4819.

MASSACHUSETTS: The ARRL Heavy Hitters Hamfest, July 20 and 21, Topsfield Fairgrounds, US Route 1, Topsfield. Indoor/outdoor flea market, food concessions, commercial exhibitors. ARRL forum, AMSAT show, CW and QSL contests, xmitter and rcver hunts, ATV demo, packet radio and musical coffeehouse (BYO instruments). Activities for non-hams. License exams, send completed Form 610, photocopy of current license and check for $4.00 payable to ARRL/VEC to: Topsfield Exams, c/o PO Box 71, Hanover, MA 02339 by June 20. Enclose SASE. Sorry no Novice exams. Free Saturday nite camping for tents and self-contained RV's. Nearby motels. Advance tickets $3.00. $4.00/door. Non-ham spouses and children free. Send check and SASE to Heavy Hitters Hamfest, PO Box 411, Waltham, MA 02254. For more information contact Russ Corkum, WA1TTV, 21 Thorndike Street, Arlington, MA 02174. *Proceeds to benefit open Waltham repeaters*, Handi-Hams and the Jimmy Fund. Talk in on 146.64 and 147.285.

THE CENTRAL OKLAHOMA COLOR OWNERS, a 278 member Radio Shack Color Computer users group, meets the second Saturday of each month at 9 AM at 10th Street and Hudson in Oklahoma City. The Club operates "COCONET", an open forum bulletin board system which can be reached at (405) 376-1494 24 hours a day, 7 days a week. The system contains COCO and FLEX operating system programs for download with no user connect fees.

WEST VIRGINIA: Wheeling Hamfest and Computer Fair, Wheeling Park, Sunday, July 21. Dealers welcome. Flea market, ARRL, AMSAT, SWOT, SMIRK booths. Family activities available at Park. Admission $3.00. To reserve space contact Jay Paulovicks, KD8GL, RD 3, Box 238, Wheeling, WV 26003. (304) 232-6796 or TSRAC, Box 240, RD 1, Adena, OH 43901 (614) 546-3930.

ILLINOIS: The Egyptian Radio Club's annual Hamfest, Sunday, June 9, 8 AM to 3 PM at ERC clubhouse and grounds. Free flea market space, approx 10', available on first come, *first served basis. Additional spaces* $5.00. Tickets $1.00 advance, $2.00/door or 3/$5. Talk in on 146.16/76 or 146/52 simplex. For information SASE to Egyptian Radio Club, PO Box 562, Granite City, IL 62040.

CQ CONTEST: VHF'ers please note! The first annual CQ World Wide VHF WPX Contest is July 20-22, 50 thru 1296 MHZ. For details, logsheets, etc., write to SCORE, PO Box 1161, Denville, NJ 07834 or to CQ Magazine. We need your entry to make this a success.

OPERATING EVENTS

"Things to do..."

Colorado Six Meter Invitational Net Contest, 0000Z July 4 through 0300Z July 5. Exhange signal report, state, name and SIN number. Send SASE and log extracts for awards to W0ETT, PO Box 6602, Denver, CO 80206 within 30 days of contest.

The Cape Fear Amateur Radio Society, of Fayetteville, NC will operate WA4YZF from 8-5 PM EDST around 7.235 MHz on *Saturday, June 15 from the 17th annual National Hollerin'* Contest in Spivey's Corner, NC. Certificate available upon receipt of your QSL along with $1.00 to help cover printing and postage. Send to Hollerin', WA4LZD, PO Box 332, Dunn, NC 28334. Allow 4 weeks for delivery.

The Tusco Amateur Radio Club, W8ZX, will operate from Fort Laurens Ohio State Memorial in conjunction with the Brigade of the American Revolution's re-enactment of 18th century military encampment 1400Z June 29 through 2200Z June 30. Special commemorative confirmation will be issued. Send legal SASE (3 IRC's for DX) and QSO info to William K. *MacNealy*, WD8LFM, RR# 1 DRH, Bolivar, OH 44612.

WEST COAST 160 Bulletin Summer SSB Contest. SSB July 13 to July 14. Time 0000 GMT 7/13/85 to 2359 GMT 7/14/85. Single operator only. Exchange RST, QTH. Class: subscribers, non subscribers. Log info: date, time, rst, OTH. Send logs to: R. Koziomkowski, 5 Watson Drive, Portsmouth, RI 02871 prior to August 31, 1985.

High Plains ARC will operate K7YPT at Historic Fort Laramie from 0000Z July 4 until 0000Z July 5. Phone — 3.850, 7.250, 14.250, 21.360, 28.550. CW — 50 kHz up from lower band edge. QSL for business SASE to: K7YPT, PO Box T, Torrington, WY 82240.

The Nazarene Amateur Radio Fellowship (NARF) will operate WA0HPW/6 to commemorate NARF's 25th anniversary during the General Assembly of the Church of the Nazarene in Anaheim, CA from June 22 to June 29. Frequencies will be 14.280, 14.305 and 21385 during daylite hours. There will be some 40 m activity. For a special QSL card, send SASE to WB6UCO, Robert Buck, 5162 W. Ave. L 12, Quartz Hill, CA 93534.

Colombian Independence Contest 1985, Saturday, July 13, 0000 GMT to Sunday, July, 14 2359 GMT. CW and phone, 1.8 thru 28. Exhange Signal report + serial number. Logs should be mailed no later than August 30, 1985 to L.C.R.A., c/o Direccion de Concursos y Diplomas, Apartado Aereo 584, Bogota, Colombia, Sur America.

The Hannibal Amateur Radio Club will again be celebrating National Tom Sawyer Days, in honor of Mark Twain's boyhood home town Hannibal, MO. Saturday, July 6 and Sunday, July 7. 1500-2100 UTC both days. Listen for W0KEM. For a certificate send large SASE and your QSL card confirming contact to Hannibal ARC, 2108 Orchard Avenue, Hannibal, MO 63401.

NEW products

IC- 3200A dual bander

ICOM has announced the IC-3200A 25-watt compact full-featured Dual Bander. With only 14 front panel controls, the IC-3200A is simple to use, yet offers these standard features: frequency coverage from 140.000 to 150.000 MHz on 2 meters and from 440.000 to 450.000 MHz on 70 cm; 5 kHz fully programmable offsets for MARS and CAP repeater operation; and 25 watts output on both bands. Its size is 5-1/2 × 2 × 8-1/2 inches. The unit also includes memory lockout and scanning capability. The price is $549.

For further information, contact ICOM America, Inc., 2380 116th Avenue, N.E., Bellevue, Washington 98004.

Circle #301 on Reader Service Card.

new Heathkit catalog

Over 400 electronic products in kit form, including the new HERO® JR. home robot, are featured in the latest Heathkit catalog.

HERO JR., the home and personal robot, is fully preprogrammed with speech output, light and sound sensors, and ultrasonic sonar, drive and steering motors, and an on-board computer system. He sings songs, plays games, recites nursery rhymes and can act as a security device and wake-up alarm.

The new GT-2218 HOTSHOT dialer is a unique one-number telephone dialer that quickly dials any telephone number up to 31 digits. It permits the use of alternate long distance service without the use of lengthy 13-digit phone and billing numbers and also allows an emergency number to be dialed by simply picking up the phone. The HOTSHOT features an easy-to-program memory that doesn't require battery backup since it is manually programmed.

Two new courses have also been added to the educational product line up: the EC-2001 Computer Fundamentals Course and the EE-1003 Analog Circuit Design Course.

For a free copy of the new Heathkit catalog, contact Heath Company, Dept. 150-510, Benton Harbor, Michigan 49022. (In Canada, contact Heath Company, 1020 Islington Avenue, Dept. 3100, Toronto, Ontario M8Z 5Z3.)

Circle #302 on Reader Service Card.

computer interface communications terminal

Amateur-Wholesale Electronics has announced its new θ-777 computer-interface terminal, featuring RTTY, bit inversion, AMTOR modes ARQ, FEC and SEL-FEC, ASCII and CW, any speed, any shift (ASCII and BAUDOT).

The θ-777 is a self-contained unit, with software, that allows reception and transmission with any computer or terminal that has RS232 or TTL I/O. The θ-777 automatically decodes signals and displays mode, speed, and polarity on the CRT. Operation is simplified by the use of 28 Bar-LEDs and LEDs including a bar-graph tuning indicator that allows precise centering of received signals.

In BAUDOT and ASCII modes, communications speed can be set from 12 to 200 baud using the modem, or 12 to 600 baud using TTL level. RS232 or TTL level data connection is 100-2400 baud (ASCII) or 45.5-200 baud (BAUDOT). Morse speed can be varied from 5 to 100 WPM in 1-WPM increments and is fully autotrack on receive.

The θ-777, which measures 2-1/2 × 9 × 10 inches, operates from a power supply of 11 to 14 volts DC.

For more information, contact Amateur Wholesale Electronics, Inc., 8817 S.W. 129 Terrace, Miami, Florida 33176.

Circle #303 on Reader Service Card.

digital FSK data modules

With packet radio and other forms of digital data communications becoming so popular, Hamtronics, Inc. has announced two new modules to its line of VHF and UHF FM transmitters, receivers, and accessories. The new modules provide for data interface with radio equipment using the popular "202" modem format (1200/2200 Hz tones) at data rates up to 1200 baud on ordinary NBFM channels. In addition to modulating and demodulating the data pulses, these modules provide transmitter keying and full handshake facilities.

The MO-202 FSK Data Modulator is a PC board module measuring only 1-7/8 × 4 inches. It automatically keys the transmitter in response to a "request to send" input from the computer, and it provides a "clear to send" handshake 25 milliseconds later, after the transmitter and receiver have had time to respond. Relative levels of the 1200 and 2200 Hz space and mark tones are equalized to compensate for the EIA pre-emphasis in the transmitter for maximum signal range. The MO-202 is only $45 in kit form, and is easy to assemble and interface with Hamtronics® and other FM transmitters or transceivers.

The DE-202 FSK Data Demodulator is a PC board module measuring only 1-1/2 × 4 inches. It can be used with any FM receiver or transceiver to detect FSK transmissions and automatically provide a "receiver carrier detect" handshake to the computer when mark or space tones are present. A special frequency compensation circuit levels the two tones coming from the receiver to allow for maximum weak signal response. The DE-202 kit is only $38.

For further information, contact Hamtronics, Inc., 65-F Moul Road, Hilton, New York 14468-9535.

Circle #116 on Reader Service Card.

frequency list

Dennis Peterson, N7CKD, Publisher of *West Coast 160 DX Bulletin* has just released the 1985 edition of the World Top Band Frequency Allocations Listing. This comprehensive list includes both CW and SSB frequency allocations for 240 of the 315 current DXCC countries, listed in the DXCC format. The list can also be used as a 160-meter DXCC check sheet. U.S. price is $5.50 (via first-class mail) and $7.50 overseas (airmail).

For more information, contact Dennis Peterson, N7CKD, 4248 A Street, Space 609, Auburn, Washington 98002.

Circle #304 on Reader Service Card.

station manager

Station Manager/Advanced is a software system that utilizes micro computer technology to fulfill the information processing needs of an Amateur Radio station. Centered around the traditional station log, Station Manager/Advanced enables the Radio Amateur to handle the routine requirements of keeping accurate station activity records with maximum efficiency as well as instantly extracting information from the station's logs that would not be readily available using conventional manual methods. It also provides the user with the capability to easily manage several concurrent logs and a comprehensive

method for logging contacts and maintaining log entries.

Station Manager/Advanced is designed to run on an IBM PC Jr., PC, or XT. Menu-driven and completely interactive, it prompts the user for all required input. Messages are displayed to indicate significant processing events and error conditions. Log entries are entered via the system keyboard through a prompt-driven input program. Once entered, log entries may be edited or deleted as required. The user may query the log entries in one of several ways. The results are displayed on the system console screen. A separate facility allows the user to print a variety of reports on the system printer. An additional feature is a special interactive update program to support QSL activities.

Station Manager/Advanced is the core of a planned series of software systems to enhance the enjoyment of your Amateur activities; to harness the power of the computer and bring your station records into the information age.

For more information, contact Omega Concepts, Inc., P. O. Box 615, Troy, Ohio 45363.

Circle #305 on Reader Service Card.

new handhelds

Two new radios are available from SANTEC: the ST-200ET (2 meters) and the ST-400ET (70 cm). Both are easy to use, thumbwheel frequency switched units designed to be compatible with accessories for handhelds you may already own. Both are backed by a comprehensive two-year extended service plan from ENCOMM.

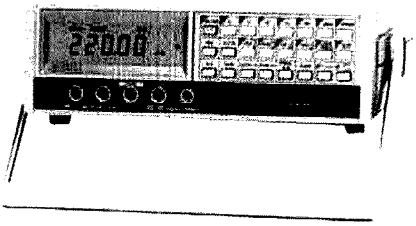
The ST-200ET is priced at $199.95 and the ST-400ET at $249.95.

For more information, contact ENCOMM, 2000 Avenue G, Suite 800, Plano, Texas 75074.

Circle #306 on Reader Service Card.

bench-style DMM

North American SOAR Corp. has announced its Model 5430 4-1/2-digit multi-function bench-style DMM. Microprocessor controlled using



SOAR Corporation's Custom LSI Chip Set, the 5430 is a 25,000-count DMM that enables users to obtain greater resolution than possible with ordinary 20,000 count units presently available.

Priced at $599, Model 5430 offers features previously unavailable in any other 4-1/2-digit DMM. It can measure DC voltage ±0.03 percent, true RMS AC voltage and current to 80 kHz, DC current and resistance from 0.01 ohms through 25 megohms. The 5430 is a dual-input frequency counter with resolution to 0.001 Hz. DC coupled to 10 MHz, it features period function as well. The input frequency can be attenuated from 1:1 to 1:1000. Temperature measurements can be made using a "K" type thermocouple ranging from −200 degrees C to 1200 degrees C with degrees F selectable. Decibel measurements from +60 dBm to −50 dBm with frequencies up to 80 kHz.

Special features include diode test, continuity beeper, data hold, peak hold with a DC acquisition time of 5 milliseconds and AC of 250 milliseconds; relative test, 3-1/2 digit select for fast survey measurements, and a comparator circuit for making Go, NO-Go tests, with comparator data output via a rear panel connector. All functions and ranges are touch-key selectable.

For further information, contact North American SOAR Corp., 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

Circle #307 on Reader Service Card.

solar panels

The ENCON Corporation has announced the release of three new photovoltaic panels: the SX-38, SX-42, and the SX-146 dual voltage-semicrystalline PV modules.

The solar cells are manufactured from semicrystalline silicon, a material developed by Solarex specifically for use in photovoltaic devices. Cells made from this material are highly efficient, stable, attractive, and inherently resistant to the "hot-spot" damage that can affect single-crystal cell under reverse-bias conditions.

Cell strings are laminated between sheets of ethylene vinyl acetate (EVA), Tedlar,® and a sheet of 1/8-inch tempered low-iron glass. This glass is self-cleaning in most climates, retains its excellent transmissivity indefinitely, and is extremely resistant to mechanical stress, including impact of hail up to 3/4-inch in diameter at terminal velocity. Furthermore, its temperature coefficient of expansion is well matched to the

cells; this matching and the stress-relieved electrical cell connections ensures excellent service even in climates with severe daily temperature ranges.

The module is framed with corrosion-resistant extruded aluminum sections with an architectural grade anodized finish. This strong, attractive framing and laminating is moisture resistant and accepts compatible mounting hardware.

The typical peaks power outputs for the new SX series modules are as follows:

SX-38	40 watts, 16.2 VDC at 2.5 amperes
SX-42	43 watts, 16.5 VDC at 2.6 amperes
SX-146	47 watts, 18 VDC at 2.6 amperes

Panels can be wired for 6 VDC with a doubling of current output.

ENCON photovoltaic systems can be used for large or small Amateur Radio projects, repeater stations, computer back-up supplies, TVRO, cellular radio, as well as residential and commercial applications.

For more information, contact Paul Denapoli, WD8AHO, at ENCON Photovoltaics, 27600 Schoolcraft, Livonia, Michigan 48150.

Circle #309 on Reader Service Card.

block downconversion accessories

LUXOR North America Corp. has introduced four new block downconversion accessories for use with its Mark 2 Block Satellite Receiver. The four DC-passing accessories include a V/H (vertical/horizontal) switch, power divider, line amplifier, and 10-dB signal attenuator.

Early STV systems used downconverters, mounted at the antenna, that sent to the receiver only a single channel from the satellite at which the antenna was aimed. LUXOR'S new "block" system permits all channels (transponder signals) on a selected satellite to be cabled into the home at once. This allows simultaneous use of multiple block receivers tuned to different channels. Thus, using a shared antenna, an unlimited number of Mark 2 Block Receivers supporting TV sets in four different locations can be individually tuned to any channel available on a given satellite.

To split the incoming satellite signal between up to four receivers, each port of LUXOR'S Model 9758 4-way DC-Passing Power Divider passes direct current from the receiver to the LNB at the antenna, allowing full flexibility in switching receivers on and off. By cascading several Model 9758 splitters, an unlimited number of receivers can be hooked up to the same antenna.

The Model 9759 DC-Passing Line Amplifier amplifies the signal 20 dB to increase antenna-to-receiver cable length or the number of receivers served by one satellite antenna. Adding one amplifier allows approximately 200 more feet of RG6 cable.

The Model 9760 DC-Passing 10-dB Signal Attenuator is designed to reduce signal level for a short run in a system that required higher signal levels elsewhere.

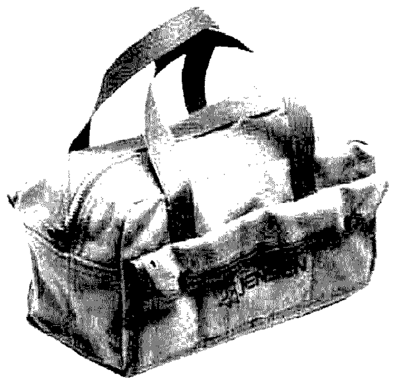
LUXOR'S new block downconversion systems can be used in homes with multiple viewers who have differing tastes in programming, in multiple-family dwellings, and in SMATV (Satellite Master Antenna Television) applications.

For more information, contact LUXOR North America Corp., 600 108th Avenue, N.E., Suite 539, Bellevue, Washington 98004.

Circle #310 on Reader Service Card.

"G.I." mechanics tool bag

The legendary "G.I." Mechanics' Tool Bags are now available from Jensen Tools in a more durable fabric (Cordura® nylon) but with the same expandability and appeal as the old canvas duck models. The bags retain the traditional green color, over-sized metal zippers, and heavy-duty web straps of the originals. Two large pockets on the outside are divided into three handy compartments each, and close with industrial strength snaps. Eight interior pouches in the main compartment serve to organize and protect small tools and parts.



For more information and a free catalog of hard-to-find tools, tool kits and cases, contact Jensen Tools Inc., 7815 S. 46th Street, Phoenix, Arizona 85040.

Circle #311 on Reader Service Card.

Kenwood TR-50 transceiver

The Kenwood TR-50 is a battery pack mobile/portable 1200 MHz FM transceiver that covers from 1260 to 1300 MHz with a 1-watt output transmitter. It features repeater offset with reverse switch, five memory channels, program-

NEW products

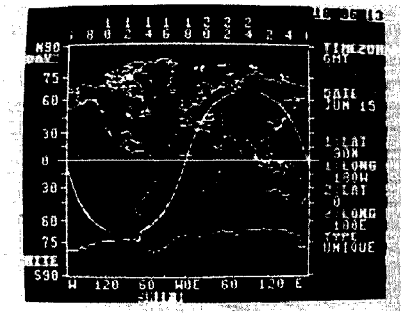
mable scanning, memory scan, a priority alert function, and RIT. "Odd splits" operation is possible using memory channel 5. The multi-function LCD panel indicates frequency, VFO A/B, repeater offset, S/RF/battery condition, and other keyboard functions.

The TR-50 comes complete with a rechargeable battery set and charger, shoulder strap, 16-key DTMF hand microphone, and an antenna on an adjustable mount. An external power cable is also included for mobile or base station supply operation.

For further information, contact Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220.

computerized DX EDGE®

Xantek has announced the availability of the DX EDGE® for the Commodore 64™ personal computer. With the new version, users of the Commodore 64™ will be able to see on-screen display of optimum QSO and reception possibilities, the Gray Line, sunrise and sunset times, and areas of the world in daylight and darkness. The display is a graphic presentation showing a detailed world map, not just a set of numbers.



The computerized DX EDGE gives all the DX advantages of the original plastic version, and additional flexibility as well. The DX EDGE finds unusual long range propagation paths by showing all the locations in the world where the sun is rising and setting at any time of year and time of day. (Excellent DX conditions often occur at these times.)

Easy to use, the DX EDGE works for all bands from 1.8 to 30 MHz. DX EDGE for the Commodore 64 is priced at $34.95. A single disk and full instructions are included.

For additional information, contact Xantek, Inc., P.O. Box 834, Madison Square Station, New York, New York 10159.

THE GUERRI REPORT

Ernie Guerri
W6MGI

predicting equipment failures

As electronic equipment becomes more complex, so too do the ways in which it can fail. This increased complexity requires greater circuit and component density; as more and more components have been squeezed into smaller spaces, the number of circuit interconnects, and hence the number of potential failure points, have also increased — by a factor of a thousand or more in just the past few years. Although we are finding ways to dramatically reduce component size and the amount of power required to perform discrete functions, the fact remains that every capacitor or resistor still has two connections, and each transistor, three . . . and so on.

Because of the small mass of modern components, shock and vibration failures are no longer the problem they once were. High circuit density now makes heat a major culprit in circuit failure. Although a modern component may have dimensions of only a few thousandths of an inch, and may dissipate only a few milliwatts, these characteristics may result in a component dissipation of several watts per square inch. Unless provision for removing the heat is made, component or device failure will occur.

An additional source of potential failure is the chemical and metallurgical processes used during fabrication. Assembling electronic circuits no longer consists of simply soldering together some tin, lead, and copper; a modern integrated circuit may require the amicable association of dozens of materials in a way that provides the opportunity for hundreds of different chemical interactions. The opportunity for one or more of these materials to contaminate nearby components is an important consideration in modern circuit failure analysis.

Because of the complexity of modern circuits, the possibility exists for thousands of combinations of potential failure mechanisms to occur in a single piece of equipment. Computers are now used to explore all of the possible relationships that could exist, and the ways in which they *could* promote failure. These techniques [known as diagnostic flow routines — Ed.] have become very sophisticated, and are deemed absolutely essential in military and space systems where no opportunity for repair may exist, and the consequences of failure can range from the merely disappointing and expensive to the catastrophic. The traditional method of expressing the likelihood of failure is called "mean time between failure" (MTBF); high-quality electronic equipment averages an MTBF of about 5000 hours.

These improvements in studying failure mechanisms and taking action to anticipate and prevent *failure mean* we can take a new transceiver out of the box, plug it in, and expect to use it for years without trouble.

new batteries promise more power, longer life

For most of us, carbon or alkaline batteries are the power source for portable equipment. "Exotic" means we spent some money and bought Nickel-Cadmium cells. However, in the world of battery developers there's some very snazzy stuff going on.

Before we examine some of the details, and their implications for Amateur Radio, let's recall some of the rules regarding the availability of energy. In general, the amount of energy we can get out of any process is a function of how active the molecules, or atoms, of the energy source can be made. Since molecular activity translates to temperature, there is a general relationship between temperature and energy release. All of the processes with which we are familiar span the range from absolute zero — where nothing happens — to fusion reactions at a few hundred million degrees C, where everything tends to come apart.

These principles apply very much to the world of chemical batteries, and in the last 15 years or so, they have inspired some dramatic developments. Three important applications that have supported these developments have included electric vehicles, space systems, and military equipment. Each

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of these applications requires large amounts of power (hundreds of kilowatts) to be delivered over sustained periods of time — i.e., several days. This means that their associated battery systems must either be very big or have high energy densities. Since big batteries are not very compatible with portable systems, high energy densities are required.

Some of the battery techniques that can provide these higher energy densities include nickel-zinc, zinc-bromine, sodium-sulphur, and lithium-metal sulphide. Nickel-zinc batteries fundamentally operate at *room* temperature and have been fabricated in forms similar to automobile batteries for electric vehicle applications. They typically have energy densities of four to five times that of comparable lead-acid devices. At the high end of the temperature scale, lithium-metal sulphide batteries operate at several hundred degrees centigrade, and can provide energy densities ten times that of the best lead-acid batteries. Although each of these battery types is commercially available for use in appropriate high-priced applications, all have proved too expensive for the operation of electric vehicles.

The most sophisticated developments in battery technology are devices using circulating fluids which depend on the chemical reaction of lithium thionyl chloride. Utilizing this technology, batteries that can deliver megawatts of power for several days have been developed. (As you might expect, you don't want to be around if you short one of these!)

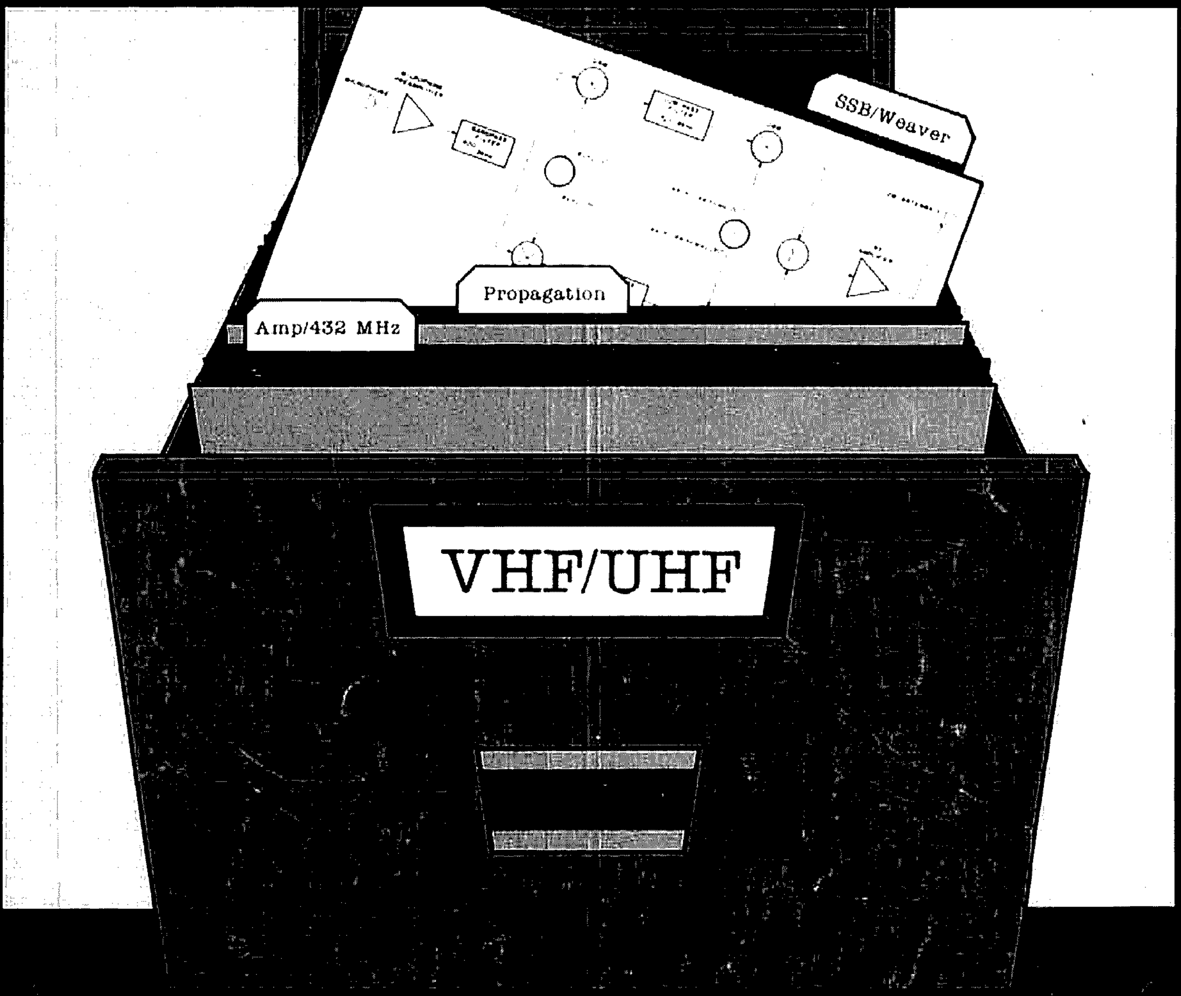
On a more practical scale, the next *few years could bring us rechargeable* lithium cells that would enable hand-held transceivers to operate satisfactorily for several months on a single charge. Bench-top transceivers with internal batteries capable of operating for several days, or weeks, are on the horizon. This would not only be a convenience for those who operate portable stations, but would also provide an important new impetus to the very function which justifies Amateur Radio — emergency operation.

ham radio

JULY 1985 / $2.50

annual **VHF/UHF** issue

ham radio magazine



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JULY 1985

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ham radio

magazine

contents

REFLECTIONS

need we be concerned?

I have a** nasty **habit. I'm *an* inveterate reader. You could almost say I'm compulsive about it. I read *everything* that comes across my desk and have a library — both at *ham radio* and at home — that for volume, rivals anything else I own. To me, it's just as enjoyable to read little tidbits here and there as it is for other people to have those little in-between-meal snacks. But a strange thing happens to you in the process of reading everything you can lay your hands on. You start to see correlations between supposedly independent facts or events.

Before you read the next paragraph, please understand that I've only begun to delve into this issue, and that what I'm about to say has potential for causing alarm. At *ham radio,* we believe that writers and editors are ethically obliged to thoroughly research every story before a drop of ink is spread. But now — given the seriousness of the news, and the credibility of the sources — I'm going to bypass that fundamental principle of responsible journalism and call your attention to a potentially widespread and significant hazard.

A small article that originally appeared in a recent issue of *Broadcasting* was picked up and retransmitted by the W5YI Report of May 15. What caught my eye was an item about a series of complaints filed by residents living near Seattle's Cougar Mountain. Their concern was with "excessive exposure to RF non-ionizing radiation resulting from 21 towers housing FM broadcast, microwave, and two-way communications antennas."

"So what," you say, "We've all heard, one time or another, about the danger of exposure to fields in excess of 10 milliwatts/cm²." The important point in W5YI's story was that "a growing number of scientific studies suggest the possibility of cell damage even *without* a recognizable rise in body temperature." Until now, I was under the impression that for damage to occur, a rise in temperature also had to occur.

Continuing on the same subject, W5YI noted another story that appeared in the April 27th issue of *Science News,* in which a Washington State epidemiologist reported "a provocative study linking death from leukemia with employment in professions that suggested possible exposures to high electric and magnetic fields." According to W5YI, the study of 546 New Zealanders identified as leukemia patients found "a significant excess of leukemias among electrical workers and radio/television repairers."

A third story cited, among hams diagnosed as having a particular form of leukemia (myeloid), a higher incidence of mortality than found among members of the general public with the same disease.

There are those who'll say that statistics can be presented to prove any contention, and maybe that's so. But I do seem to recall that the Soviets have a much lower acceptable limit for non-ionizing radiation: approximately 100 *micro*watts/cm².

Perhaps this would be a good time to re-examine the amount of radiated power from that unbalanced transmission line, single-wire feed antenna, or close-by antenna. Using a backwards argument, maybe there's even some good sense in placing your antenna higher, on a taller tower. Doing this certainly won't hurt its performance.

Let me assure you that this isn't just a thinly veiled attempt on my part to influence you all to go QRP so my puny signal will be more competitive . . . it's an attempt to call your attention to a serious matter that will be discussed more fully as more information becomes available.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

GREATLY EXPANDED NOVICE PRIVILEGES WERE PROPOSED by ARRL's Executive Committee at its May meeting in Rochester. Basically, the committee recommended giving Novices both phone and digital privileges on three bands, two of them UHF. Perhaps the most significant element of the proposal is the addition of both phone and digital communications privileges on 10 meters. There the Novice band would become 28.1-28.5 MHz, with the bottom 200 kHz for RTTY, AMTOR, and packet as well as CW and the 28.3-28.5 MHz slot SSB and CW only (to preclude use of converted AM CB radios). Under the present rules Technicians would also gain 10 meter privileges along with Novices; power for both would remain 200 watts out but that limit wouldn't apply to others. One important problem the expansion would cause is with the 10-meter beacon band, now 28.2-28.3 MHz; some possible alternatives for beacon operators were also discussed.

On UHF The Committee Recommended Giving Novices Full Privileges on the entire 220 MHz band and 1246-1260 MHz. They'd be permitted to use but not put up repeaters. Power output for Novices would be limited to 25 watts on the 220-MHz band, and 5 watts on 23 cm. The new privileges would require some expansion of the Novice exam questions; already licensed Novices would, of course, be grandfathered into the new modes and bands.

At Presstime This Is Still Only An Executive Committee Proposal for the League Board of Directors to review, then adopt or reject. However, the concept of expanding Novice privileges as a means of making the entry level license more attractive has seen increasing support inside as well as outside the League for some time. In addition, FCC Special Services Chief Ray Kowalski indicated at Dayton that the Commission was also looking with favor toward a more attractive Novice license package.

If The Directors Do Vote In Favor Of Increasing Novice Privileges, look for the ARRL to file a Petition for Rule Making to the Commission very soon. Since the future of the 220 MHz band is still under a cloud, the League's proposal will probably be worded in such a way that the FCC can leave that band out of any resulting NPRM if it so chooses.

ALL VECS ARE INVITED TO GETTYSBURG AUGUST 8, when the FCC will host a familiarization meeting for them. Purpose of the all-day session is to permit the VEC representatives and the FCC people they work with to meet each other face to face, so both can better appreciate the problems each has with the program and the licensing system. Particular attention is planned for paperwork errors; though some VECs are very good, a survey of April applications showed almost two-thirds of one VEC's submissions had significant errors! Timeliness is still another problem that's to be addressed at the meeting, with some VECs chronically late in submitting their exam session results.

THE SPACE SHUTTLE'S PRIME AMATEUR BAND DOWNLINK FREQUENCY will be 145.55 MHz, and at presstime launch was still scheduled for July 15. Amateur operation could occur as early as Mission Day Two, though expectations are that little, if any, of WØORE's or W4NYZ's time on the air will be devoted to unscheduled two-way QSOs. They will be doing a good deal of work by prearrangement with various schools and clubs, however, and hope to provide live or pre-recorded SSTV downlink signals during periods when they can't be on themselves.

RTTY, FAX, AND SSTV WERE ALL AUTHORIZED ON 160 METERS effective June 17. In its Report and Order on PR Docket 84-959, the FCC decided to permit the use of all three modes across the entire band, but cautioned that introduction of the new modes does not temper the possible reallocation of 1900-2000 kHz to radiolocation in Docket 84-874.

THE INDUSTRY GROUP THAT'S BEEN WORKING ON AMATEUR RADIO'S FUTURE had its second meeting in Dayton, the Thursday night before the Hamvention. About 40 industry representatives attended and heard the task force leaders report considerable progress on various promotional efforts. Travis Brann, WA5RGU, of Kantronics, succeeded Mike Lamb of AEA as de facto group chairman for the next quarter; Joe Schroeder, W9JUV, will continue to act as secretary/treasurer for the time being. A delegation from the group is scheduled to meet with Senator Barry Goldwater, K7UGA, for a briefing in early June.

The First Attempt At Implementing One Of The Group's Proposals Appears to have met with some success at the Rochester Hamfest in May. A good number of the free tickets sent to area junior high and high school teachers for distribution to interested students were used, and special booths aimed at entry level prospects were reported to be very popular.

The Average Age Of New Amateur Licensees In April was 36, the FCC determined after analyzing the approximately 2200 applications from never-before licensed applicants it received that month. The oldest was 82, the youngest 7, and the median age 35.

SOME FOUNDATIONS FOR A NATIONAL REPEATER COORDINATION SYSTEM were laid during the course of several well-attended meetings at Dayton. The first, organized by W8JRL (Ohio) and WB8UPM (Michigan), discussed the relative merits of 15 vs 20 kHz channel spacing, while the second was on the FCC's national repeater coordinator proposal, PR Docket 85-22. At that meeting ARRL Hudson Division Vice Director WA2DHF reported tentative League agreement to fund a computer system and incoming WATS line for coordinator use, and to publish a coordinator's newsletter. At the third session the consensus seemed to favor a "confederation" of area coordinators rather than a single national coordinator.

Frequency Spectrum Chart

NOT AVAILABLE — CW

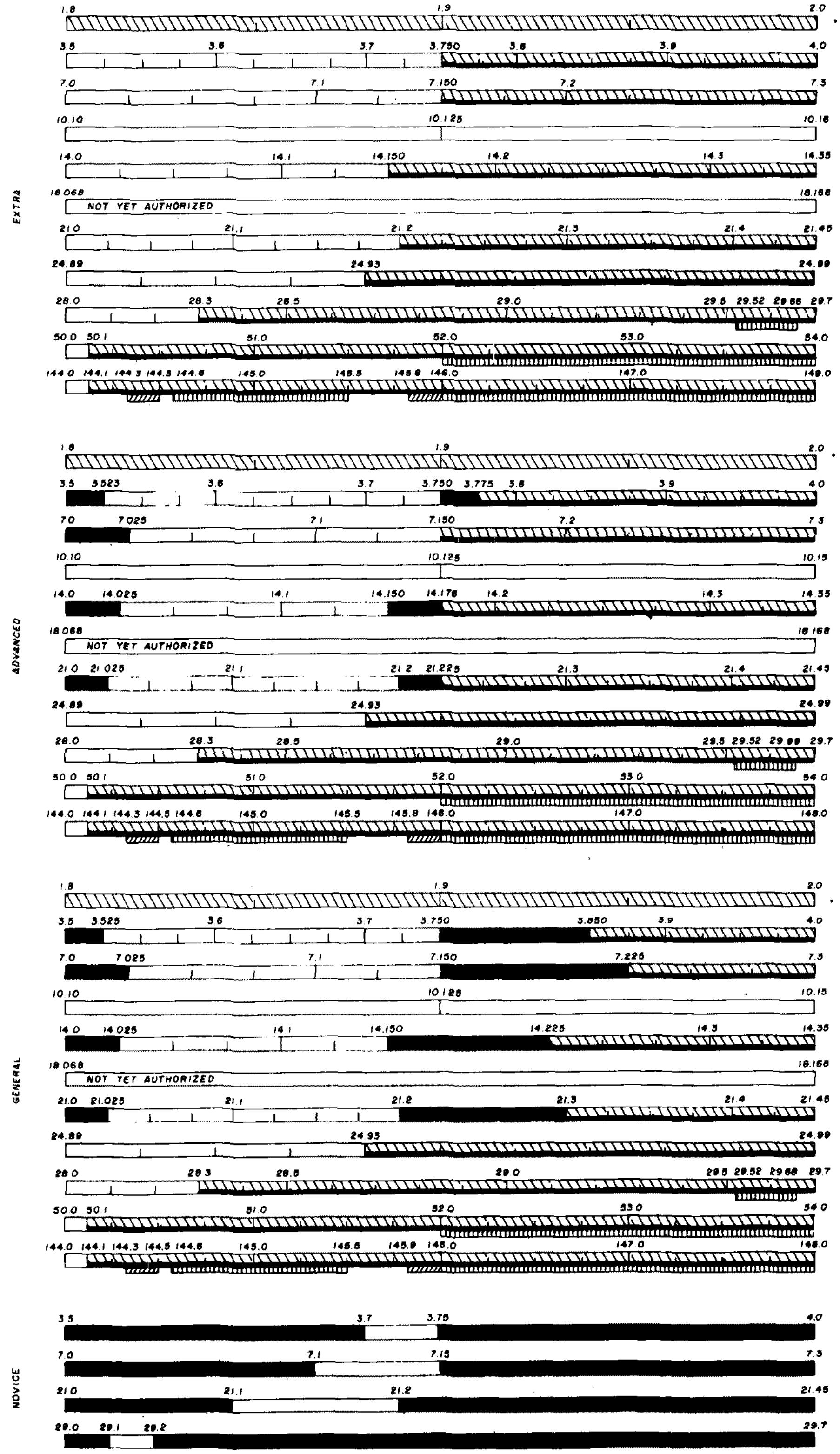
REPEATER — PHONE

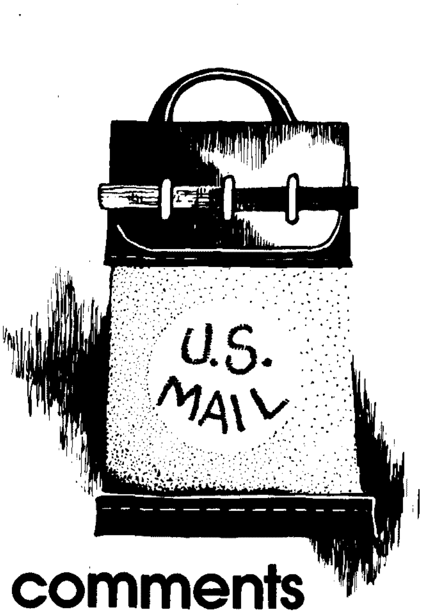
OSCAR — PHONE SSTV

TECHNICIAN LICENSEES have privileges of all Novice frequencies plus General frequencies above 50 MHz.

CW permitted on all frequencies available to that licensee.

*effective June 17, 1985 RTTY, FAX, and SSTV modes also available.





comments

active antenna

Dear HR:

The article in May, 1985, issue of *ham radio* on active antennas ("Active Antenna Covers 0.5-30 MHz," by Peter Bertini, K1ZJH) was interesting but contained no evaluative data. Your readers might be interested in knowing that an evaluation of S/N performance of active antennas was published by Radjy & Hansen in the March, 1979, *IEEE Transactions on Antennas and Propagation*, Vol, AP-27, pages 259-261. S/N degradation occurs over the upper part of the HF band where external noise reduced by the impedance ratio is less than preamp noise. This degradation is aggravated by large design bandwidth since that requires a large impedance ratio.

Robert C. Hansen
Tarzana, California

On-the-air support

Dear HR:

I agree that we need an aggressive Junior High ham radio program — but we need actual on-the-air support just as badly, or maybe even more so.

A year and a half ago I was working with a 14-year old and a 13-year old. Those kids picked up code so fast I couldn't believe it. They were doing great on element 2. They weren't interested in Novice tickets — they wanted General Class licenses, and with some on-the-air support they would have made it.

I loaned them each a transceiver (Sans finals — hi!) so they could look around the bands for something of interest to them. I looked for contacts I thought would be interesting to them. They worked many contacts on my rig looking for teenage support. They didn't call it that, but follow-up conversations made it obvious they couldn't find anyone on the air that shared any interests with them.

Needless to say, we — the ham world — lost two bright youngsters for what I believe was lack of interesting QSOs for two teenagers. . . . Both of them turned to computers. About 300,000 people live in our metro area, so they now find lots in common with other kids on their computer nets. They're also interested in interfacing music synthesizers with their computers.

I've tried to get them back into Amateur Radio, but I can't compete with their age group. . . .

We're friends, and the three of us play golf together. I have a feeling that when they can beat me at golf, they'll be off to greater challenges — but for now, we share that interest.

Any ideas on how to hold their interest? The neighborhood is full of youngsters coming into their teens, and I'm willing to try again if all of us will try to share their interests when they're on the air.

Ken Uthus, KT7E
Nine-Mile Falls, Washington

CQ DXRC?

Dear HR:

I was very pleased to see your comments on courteous practices and the "rubber stamp" QSO ("Reflections," March and April, 1985). What made me lose interest years ago and still threatens is the fact that in *any* DX QSO, even from New England to Europe, which is a very solid circuit on almost any band with favorable propagation, you rarely find a conversation longer than a minute. Now I don't knock awards and contests — to each his own — but to me the fact that I have the capability, with my rig, and knowledge to exchange meaningful information with individuals in foreign countries is the most exciting and interesting aspect of Amateur Radio.

I don't ever wish to monopolize a DX station's attention when others are calling. But there have been many, many times I have worked a DX station who cut it very short only to call "CQ DX USA" two or three times before getting another QSO. This is frustrating to me, because my own definition of "rare DX" is the guy, wherever he is, who's willing to ragchew for a few minutes.

I suggest this happens because there is so much contest and certificate operation that the short QSO habit develops. Is there a way we can encourage both domestic and foreign stations to chew the fat for a few minutes when there aren't any pile-ups happening? Perhaps a new general call, like "CQ DXRC," which would imply that the caller isn't looking for the rare prefix, but rather a little international fellowship instead.

Anyway, I sure would like to see some movement in the direction of some of us getting to know each other, and I believe it can be done without jeopardizing the DX operators who are looking for their country totals. So often, in routine conditions, there are lots of Europeans working lots of Americans and nobody is finding out anything except "579 name is Bob tks 73 gb. . . . "

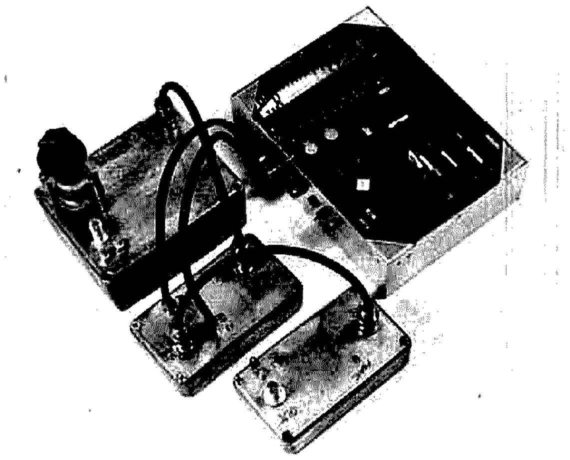
David Lewis, KA1KFC
(ex WA2ZQU)

OSCAR 10

Dear HR:

Many thanks for publishing "A PSK Telemetry Demodulator for OSCAR 10" (April, 1985, page 50). Why, I wonder, has it taken so long for this to appear in an Amateur publication?

Stephen E. Bach, AA4B
Scottsville, Virginia



2-meter transmitter uses Weaver modulation

Try the "third method" of SSB generation

Imagine a 2-meter SSB transmitter that contains no crystal filters, no IF amps, no heterodyne oscillators, no BFOs, and no broadband audio quadrature generator, either.

Impossible? *No.* The scheme described herein is based on a little-known technique usually called the "third method" of SSB generation, which I prefer to call the "Weaver Modulator," in honor of D.K. Weaver, its apparent inventor. First discussed in 1956, the technique has rarely been seen in the commercial or Amateur press.[1]

The purpose of this project was to demonstrate that the Weaver modulation technique could be easily and inexpensively applied to direct conversion sideband generation at VHF frequencies. It was not my intention to build a full-function rig, but merely to experiment with the architecture; therefore, the design does not include any T/R switching, ALC circuitry, or digital frequency display. Intrepid homebrewers can easily add these functions themselves.

Despite the fact the "filter" technique of SSB generation has been almost universally adopted for Amateur and commercial design, the Weaver technique offers the following advantages:[2,3,4]

- Much of the circuitry operates at audio frequencies, where layout is relatively non-critical. Components for these applications are inexpensive and easy to obtain.
- There is only one RF oscillator, and it operates at the center of the transmitted output passband rather than being offset by an IF frequency. The oscillator may be tested with ordinary Amateur equipment; a 2-meter receiver can be used as a detector. Also, a conventional frequency counter can be used as a digital frequency readout, since there are no BFO or IF offsets to account for.
- All of the mixers operate on baseband signals. The absence of heterodyne techniques mean that there are (theoretically, at least) no images or spurs. Any out-of-band radiation is a result of mixer and amplifier non-linearities, and not a result of any inherent limitations of the conversion scheme.
- Unlike the "phasing" technique, the Weaver modulator does not depend upon accurate phasing or balancing to achieve good control of the transmitter bandwidth. Phase and balance errors cause degradation of the audio quality only, not out-of-band components.
- No expensive or hard-to-find crystal filter is necessary. For the most part, no unusual components are required; the average junkbox probably contains most of the components needed for the design.

By Norm Bernstein, N1COX, 24 Foxfire Drive, Sharon, Massachusetts 02067

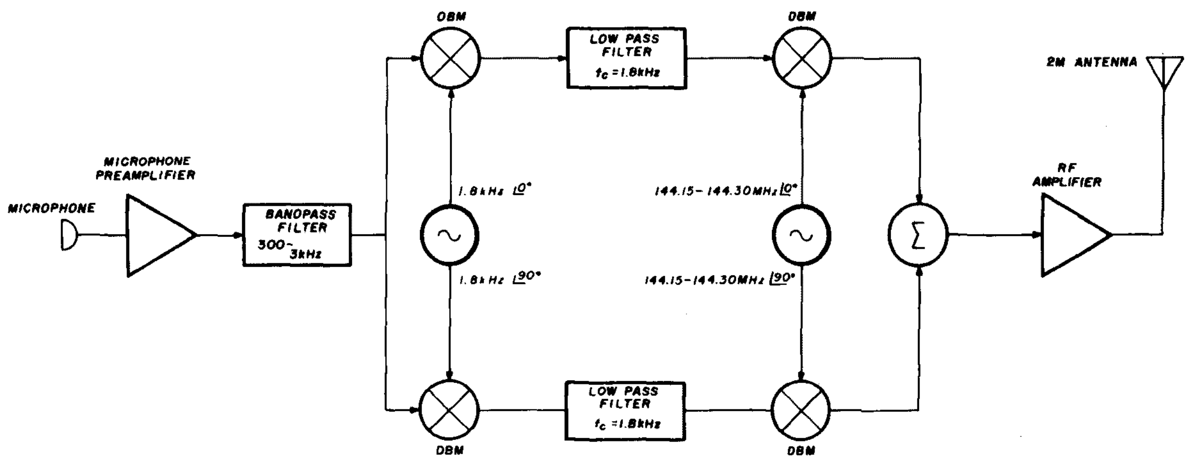


fig. 1. Block diagram of a single sideband transmitter using the Weaver modulation technique.

The only significant disadvantage of the Weaver modulator technique is local oscillator suppression. In a conventional SSB transmitter, any LO leakage is located at the position of the suppressed carrier, and a filter type receiver would normally be tuned to zero beat this signal for proper reception. In the Weaver modulator, however, the LO is set exactly in the middle of the transmitted passband, and good RF mixer balance is essential to avoid an unpleasant "whistle" on the transmitted signal. Fortunately, commercial DBMs have excellent balance characteristics, and the carrier leakage can be dealt with successfully. (Interestingly, this would not be a problem if the intended receiver used the Weaver technique as a demodulator, since the receiver's local oscillator would then zero beat with the transmitter's leaky LO signal, rendering the leakage inaudible.)

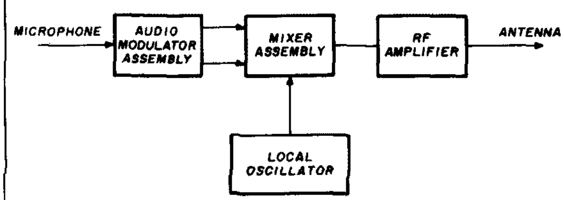


fig. 2. An audio modulator, mixer assembly, local oscillator, and post mixer amplifier comprise the modular construction design approach.

circuit description

Figure 1 shows a block diagram of the basic technique. The signal from the microphone is amplified and filtered for the normal 300 Hz to 3000 Hz communications bandwidth. It is then applied to a pair of double balanced modulators; the modulators are driven from an audio frequency local oscillator whose outputs are in quadrature. The AF local oscillator runs at 1.8 kHz, which is the center of the audio passband.

The outputs of the DBMs are then fed to a pair of low-pass filters, each with a cutoff frequency of approximately 1.8 kHz. These filters establish the basic transmitted bandwidth and are analogous to the crystal filter found in conventional rigs.

The outputs of the filters are then sent to another pair of double balanced mixers; these mixers are driven from a quadrature local oscillator operating at the desired RF frequency. The outputs of the mixers are then summed, with the resultant output being a single sideband signal. The signal may then be amplified in a conventional manner before being fed to the antenna.

Selection of the upper or lower sideband can be made by switching the phases of either of the local oscillators or by swapping the outputs of either pair of double balanced mixers.

While the actual technique might be difficult to understand, its mathematics are relatively simple. Rather than attempt a complete description of the mathematics at this point, I recommend that interested readers consult the references listed at the end of this article, especially the original paper by Weaver.

designing the prototype

To minimize leakage effects and simplify testing, I decided to *split* the design of the prototype into four functional blocks: the audio modulator (containing the microphone preamplifier, the audio double balanced mixers, the filters, and the AF local oscillator generating circuitry), the local oscillator, the RF mixer assembly, and the post-mixer amplifier. The audio modulator was housed in an aluminum chassis box,

and the remaining three modules were constructed in separate die-cast boxes, using BNC connectors for signals and feedthroughs for DC power. Both **fig. 2** and the photo show the interconnection of the four modules.

The audio modulator assembly performs all the baseband signal processing, producing an output suitable for driving the RF mixers directly. This module has the most complex circuit of the four, but is the easiest to build because the layout is not critical; I used a conventional punched board with sockets for the ICs and point-to-point wiring.

The microphone input is connected to a wideband gain stage (**fig. 3**) in order to bring the audio signal up to the nominal working level (2 to 3 volts p-p). The signal is then fed to a highpass filter in cascade with a low-pass filter. These filters are implemented as third-order Sallen and Key types with cutoff frequencies of 300 Hz and 3 kHz, respectively.

The signal is now split into two paths. Each path consists of a double balanced mixer, followed by a relatively sharp low-pass filter, followed by a buffer stage and 50-ohm pad designed to deliver approximately 0 dBm to the mixers.

The double balanced mixers are implemented with a series/shunt switch (1/2 of a CD4016 CMOS switch) and an op amp configured as an "ihvert/non-invert" stage. This type of mixer exhibits good linearity and balance at audio frequencies, but has strong spurious response at harmonics of the local oscillator frequency; this is why the microphone preamplifier is followed by a relatively sharp bandpass filter.

The signals from the mixers are then routed to the low-pass filters. The filter characteristic is important

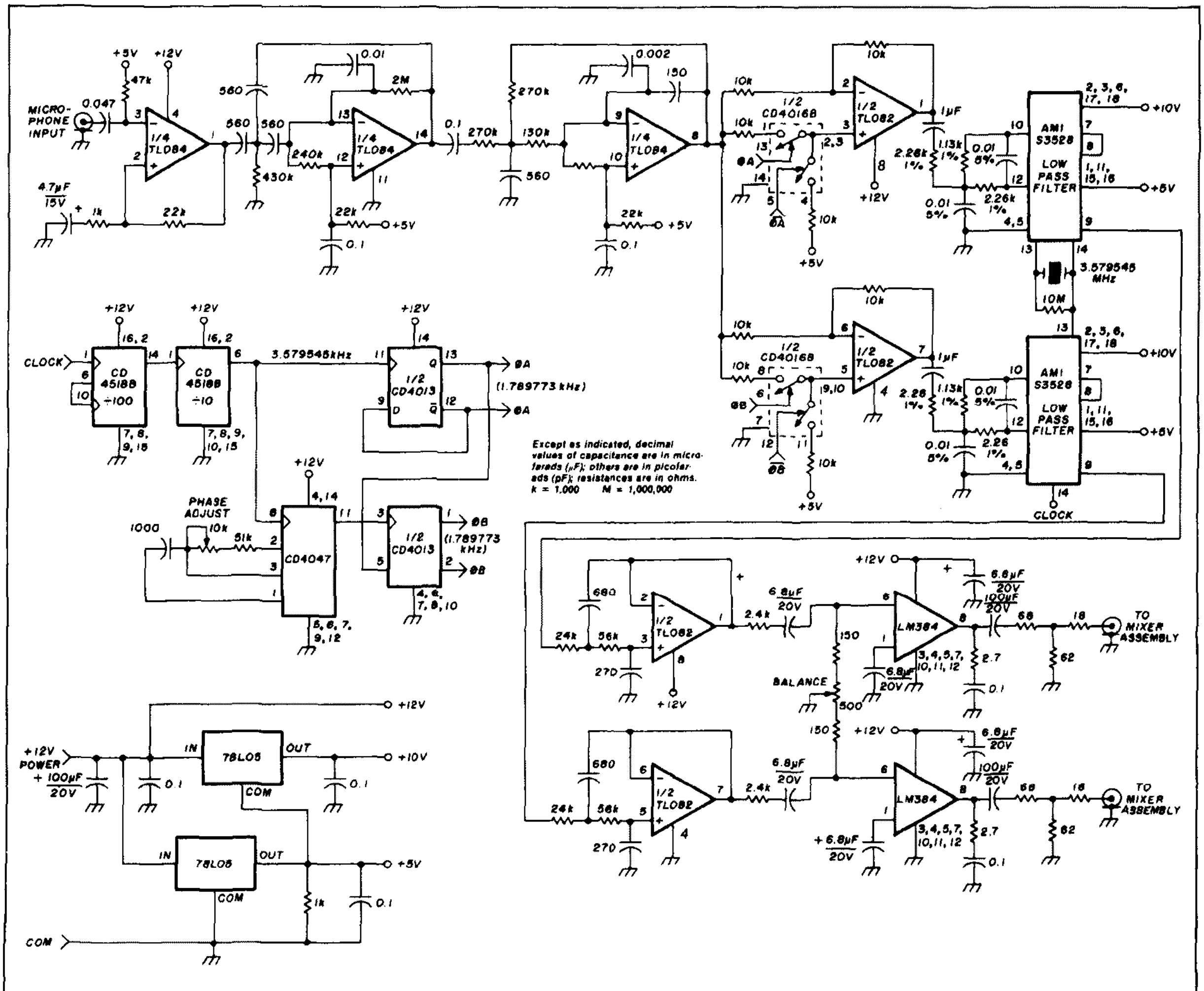
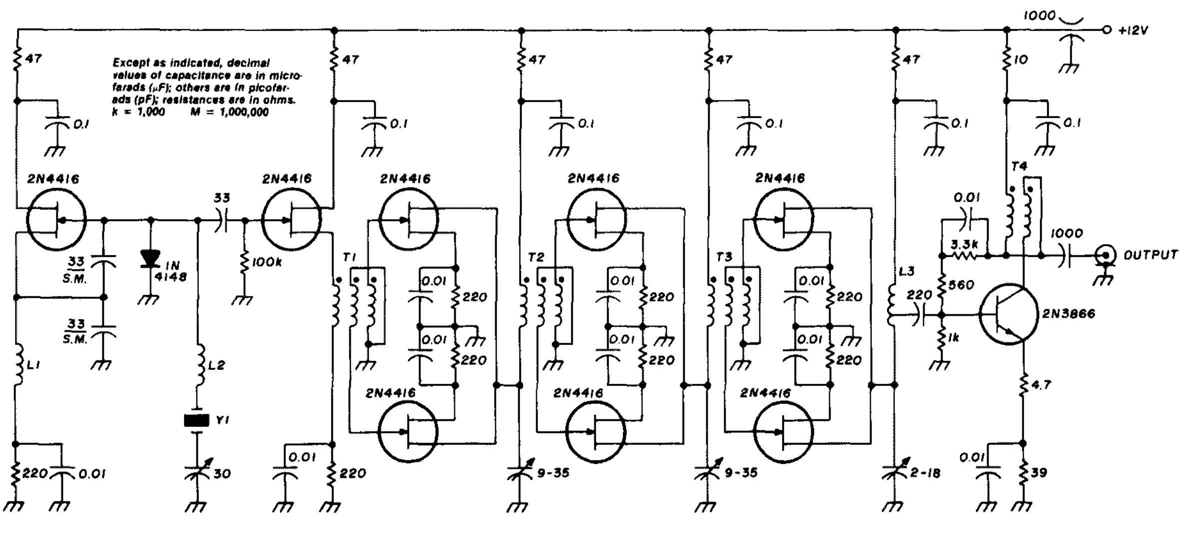


fig. 3. The audio modulator assembly consists of a microphone preamplifier and bandpass filter, a pair of double balanced mixers, and the switched capacitor filters. Also included in this assembly are the 1.8 kHz quadrature local oscillator and a pair of amplifiers for driving the RF mixers.



L1	33 turns No. 24 enameled, FT50-61 core
L2	35 turns No. 24 enameled, T50-6 core
L3	4 turns No. 18 tinned, 1/4" ID × 1/2" long tap 1/2 turn from cold end
T1	12 turns No. 30 enameled, trifilar on FT50-61 core
T2	11 turns No. 30 enameled, trifilar on T37-6 core
T3	7 turns No. 30 enameled, trifilar on T37-6 core
T4	5 turns No. 24 enameled, bifilar on FT37-61 core
Y1	18.025 MHz, parallel resonant, 20 pF load capacitance

fig. 4. The local oscillator is a VXO operating at 18.025 MHz followed by three push-pull doublers and an amplifier. The output is 144.150 to 144.300 MHz at about +17 dBm.

because it affects the degree of unwanted sideband suppression and establishes the bandwidth of the transmitted signal in much the same way that the crystal filter does in a conventional rig. What is needed here is a high order elliptical low-pass filter with good stopband performance. One additional requirement is that the filters in each of the two signal paths have closely matched amplitude and phase characteristics; any significant mismatch here will affect the unwanted sideband suppression (which, in a Weaver modulator, results in a degradation of the audio quality).

While the filters could have been built using conventional LC techniques, I decided to use a pair of switched capacitor filter ICs. This device the S3528 from American Microsystems, Inc.,* is a seventh-order elliptical low-pass filter with a programmable cutoff frequency and better than 50 dB worth of stopband suppression. A pair of these devices is significantly smaller than corresponding passive LC filters and are "tweak free" — i.e., they require no adjustment whatsoever and are inherently well matched. A minor disadvantage to switched capacitor filters is that they require some additional filtering at the input and some filtering at the output to remove the residual clock component from the signal, but this was not difficult to accomplish.

After the filters, the two audio signals go to a balancing network followed by a pair of LM384 driver amplifiers. These amplifiers are power devices capable of driving the 50-ohm pads used to reduce the signal to approximately 0 dBm, a level appropriate for the IF ports of the RF mixers. The heavy attentuation also helps to insure that the mixers see a broadband resistive termination at their IF ports, which is important for proper mixer operation.

The switched capacitor filters contain their own oscillator, which is based on a standard 3.579545 MHz colorburst crystal. This clock signal is divided by 1000 and applied to a pair of flip flops, one of which is delayed by an adjustable one-shot to create a 90-degree phase lag. The output of these two flip-flops is an adjustable quadrature signal operating at 1.789773 kHz, which is close enough to the design value of 1.8 kHz for suitable operation. Feedback from the non-delayed flip-flop is employed to insure consistent phasing at startup; without such feedback, the

*American Microsystems, Inc., a division of Gould, Inc., 3800 Homestead Road, Santa Clara, California 95051.

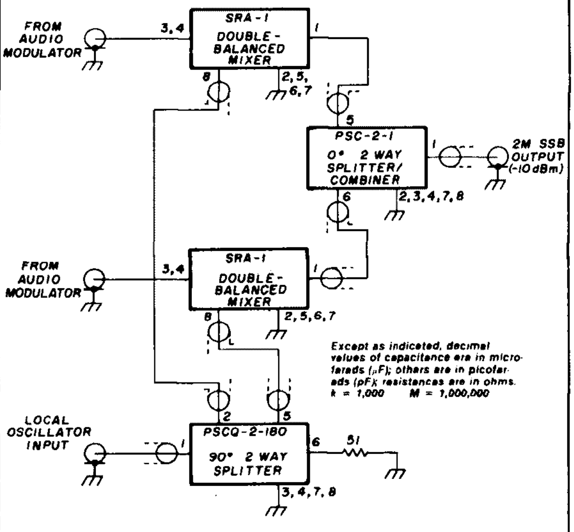


fig. 5. The mixer assembly contains two SRA-1 double balanced mixers, a 90-degree 2-way hybrid (which is used to create the quadrature local oscillator signals for the mixers), and 0-degree, 2-way hybrid, which sums the two mixer outputs into a single SSB signal.

transmitter would choose upper or lower sideband at random!

The AF local oscillator quadrature could have been generated with perfect accuracy through the use of flip-flops alone, but I used the one-shot delay in order to allow for some adjustment range; it can be shown mathematically that small phase errors in the RF mixer assembly can be cancelled by introducing an equal but opposite phase error into the system at the AF mixers.

The RF local oscillator (fig. 4) is a VXO running at a nominal frequency of 18.025 MHz, followed by three doubler stages and a buffer stage. This design is simple to build, adequately stable, and provides for enough tuning range to cover most of the portion of the 2-meter band commonly used for terrestrial communications. My version covers 144.150 MHz to 144.300 MHz; it is possible to obtain a wider coverage, but tuning ranges in excess of 0.1 percent of the nominal output frequency will result in reduced stability.

Although no frequency indicator was constructed for this experimental rig, it would be relatively easy to build one because the oscillator runs at the transmitted frequency; there are no IF or BFO offsets to account for. A general-purpose frequency counter capable of operation at 2 meters can also be employed.

The RF mixer module (fig. 5) consists of a pair of SRA-1 mixers whose local oscillator inputs are driven in quadrature, and whose RF outputs are summed into a single output. The local oscillator drive is obtained from a commercial quadrature hybrid, in this case the PSCQ-2-180 from Mini-Circuits Labs.* The summation of the RF outputs is accomplished with a hybrid combiner (model PSC-2-1). The two audio drive signals are connected directly to the IF ports.

A post mixer amplifier is used to provide 30 dB gain to the −10 dBm 2-meter SSB signal output of the mixer assembly. This results in a signal of about 100 milliwatts, which is sufficient for on-the-air testing. This amplifier (**fig. 6**) is a three-stage device with a grounded gate FET followed by two broadband bipolar class A stages. Because of the relatively low power, I did not incorporate any further filtering of the signal; more would undoubtedly be incorporated, however, in a practical design.

test results

This experimental rig was tested on the air in order to get some subjective feedback on the audio quality. The estimated output power was 50 to 100 milliwatts, too small to be accurately measured on any of my test gear. My first QSO was with W1VDI in Providence, Rhode Island, about 30 miles from my QTH. I received a Q5 report.

Listeners generally reported that the audio quality was essentially equivalent to that of my regular 2-meter SSB rig (an ICOM 251A); minor differences in tone quality were attributed to the use of a different microphone. None of the test participants reported any trace of carrier leakage on the signal, which indicates that the carrier balance of the mixer assembly is adequate.

The only negative comment from the test participants was that there was a brief (2 to 3-second) period immediately after I keyed the transmitter each time, during which some traces of carrier could be heard; this effect "died out" within a few seconds. This was found to be caused by DC bias level settling in the audio modulator section; it can be avoided by maintaining continuous power to the audio modulator rather than attempting to switch the modulator power when the transmitter is keyed.

The transmitter was later examined with an RF spectrum analyzer and found to exhibit a carrier suppression of better than 45 dB, with the reverse sideband component down at least 30 dB from maximum output.

the Weaver technique as a receiver

The Weaver technique is bilateral. If all of the elements can be constructed to operate bilaterally (**fig. 1**), then the system can be used to demodulate SSB signals as well as generate them. Although I have not had a chance to experiment with receive applications, it seems to me that many of the advantages of this technique apply in a demodulation system as well. Images and spurs would be far less of a problem than

*Mini-Circuits Labs, Inc. 2625 East 14th Street, Brooklyn, New York 11235.

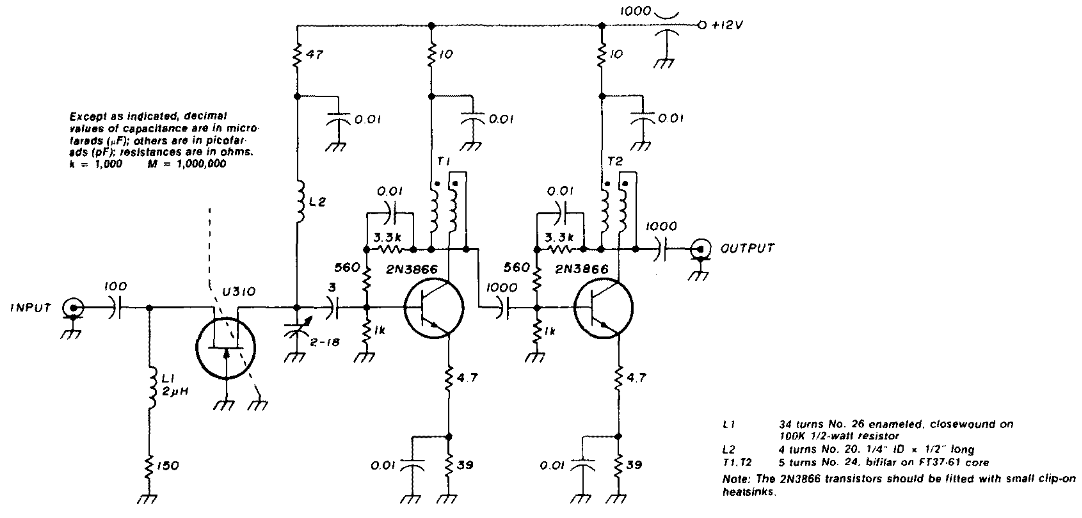


fig. 6. The post mixer amplifier provides about +30 dB of gain at an output of approximately 100 milliwatts.

in conventional heterodyne architectures. Dynamic range should be quite good, since conversion and demodulation occur in the first stage without the need for IF amplifiers, which can overload. All of the gain (with the exception, perhaps, of some RF amplification before the first mixer pair) would be accomplished at audio frequencies, where recent advances in IC processing techniques make low noise audio amplification relatively easy.

low cost variations

It should be possible to reduce the cost of this design by substituting components in the RF mixer assembly and the audio filters. The hybrid combiner and 90-degree splitter could perhaps be replaced by Wilkinson dividers (made from two 1/4-wavelength sections of 75-ohm cable, joined at one end) and a 1/4-wavelength section of 50-ohm cable for the phase delay. The cable scheme would probably have enough bandwidth and accuracy for 2-meter SSB operation, especially in view of the relatively narrow bandwidth popularly used on 2-meter SSB. Precise measurement of the cable lengths would not be necessary, since small amounts of phase error can be "tuned out" with the phase adjustment in the audio modulator section.

The audio filters need not be quite as sophisticated as the ones used in the prototype design; the switched capacitor filters could be replaced with equivalent LC designs. The differential phase performance of the two filters is important, however, for good reverse sideband suppression; it will therefore be necessary to measure the component tolerances of the Ls and Cs quite carefully.

special consideration

Any practical application of the Weaver modulator will require some special design consideration. For example, when using the prototype transmitter in conjunction with a conventional "filter type" receiver, leakage from the transmitter's local oscillator would overload the receiver front end during reception. One way to minimize this problem would be to disable the VXO multiplier stages during receive. In the interests of stability, however, it would not be advisable to key the VXO itself.

acknowledgements

I would like to thank W1VDI, K1TOZ, and WA9WTK/1, all of whom participated in the on-the-air testing phase of this project. I would also like to thank Ed Wetherhold, W3NQN, for his advice on the subject of audio filters.

references

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2. Howard F. Wright, "The Third Method of SSB," *QST*, September, 1957, pages 11-15.
3. J.F.H. Aspinwall, "The Third Method — A New System of SSB Generation," *Wireless World*, January, 1959, pages 39-43.
4. Herbert Krauss, et al, *Solid State Radio Engineering*, John Wiley and Sons, 1980, pages 233-234.
5. Joseph Sansone, "Get High-Q in Active Bandpass Filters with a Quadrature Modulation Scheme," *Electronic Design*, November 8, 1978, pages 124-127.

ham radio

VCO tunes from 1800-2600 MHz

New to Microwaves? Build and test this simple VCO

I have built antennas, preamplifiers, filters, mixers, and oscillators for microwave frequencies from 23 to 4.8 cm. All performed as expected except for the oscillators, which required multiplication of the fundamental for the 13-cm band. Some required considerably more multiplication because the fundamental frequency was as low as 15 MHz. The oscillator/multipliers were complicated — difficult to build and even more difficult to test. They produced fundamental output, multiplied output at undesired harmonics, and often spurious output. To avoid these problems, I decided to build a fundamental voltage controlled oscillator (VCO) that operated directly on the 13-cm band.

This simple VCO was easy to build and test. The tuning range, power output (2 to 15 mW), phase noise, and stability of the VCO were acceptable for many applications.

The parts for the VCO can be purchased from dealers or mail order suppliers for a total of less than $20. If all the parts for the VCO are available, including the etched PC board and test equipment described below, the VCO can be built and tested in an hour, using standard tools. A mounted low-power magnifying glass is a helpful accessory.

I built a line stretcher and microwave detector for testing the VCO, using the line stretcher to determine the approximate frequency and the detector (with a milliammeter) to measure the power output.

I've used the VCO as an FM transmitter and as a local oscillator for a converter for broadband (greater than 100 kHz) communication. Since the VCO is remotely tunable, it can be conveniently mounted at the antenna feed, thus eliminating line losses for either transmission or reception.

background

While searching for a suitable circuit for a fundamental 13-cm oscillator, I found a very simple WWII era oscillator circuit in the 1949 *Radio Amateur's Handbook*. This circuit used two 6J6 tubes with a transmission line connected to the plates of the two tubes. Oscillation was at 420 MHz. The only way the 6J6 could be made to oscillate at such a high frequency was to divide the shunt capacitance load on the transmission line by 2, using 2 tubes. The cathodes and plates of the tubes were isolated from RF ground with RF chokes, and the grid grounded through a bias resistor. Oscillation occurred because of coupling between the two sides of the common plate transmission line.

I built this same circuit on a PC board, but used two microwave transistors in place of the 6J6 tubes and spiral wound carbon film resistors as RF chokes. Using transistors in place of the tubes allowed tuning over a wide frequency range since the junction capacitance of the transistors is easily varied by changing the base to emitter current. **Figure 1** shows the circuit for this simple VCO.

Capacitance of a dual 6J6 from grid to plate is about 3 to 4 pF. Bipolar microwave transistors available today have C_{OB} of 0.4 pF. This should allow oscillation at frequencies as high as 6 GHz with a significant portion of a transmission line external to the transistors. This VCO circuit actually has the base grounded so capacitance in the transistor is higher than 0.4 pF.

Any transistor with a C_{OB} less than 0.5 pF, a DC Beta greater than 25, and F_T greater than 4 GHz would probably operate properly in this circuit. Transistors such as MRF901, BFR90, HXTR-61, and HP 35821B could be used, although the maximum frequency of operation might be lower.

I used the NE21935 in the VCO because it costs less than $7.00, is bipolar, has a convenient package form with two emitter leads, is very small, and has excellent C_{OB} and F_T specifications. The NE21935s are available

By Hans M. Roensch, Jr., W0DTV, R.R. #1, Box 156B, Brookfield, Missouri 64628

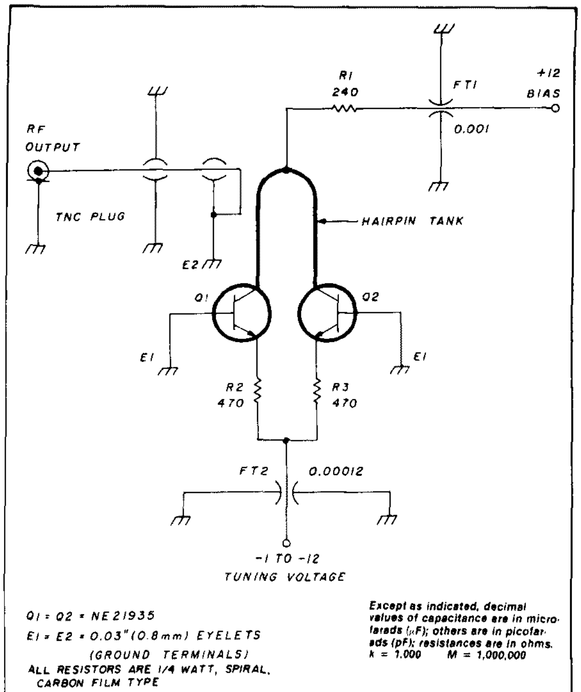

fig. 1. Microwave voltage controlled oscillator circuit diagram.

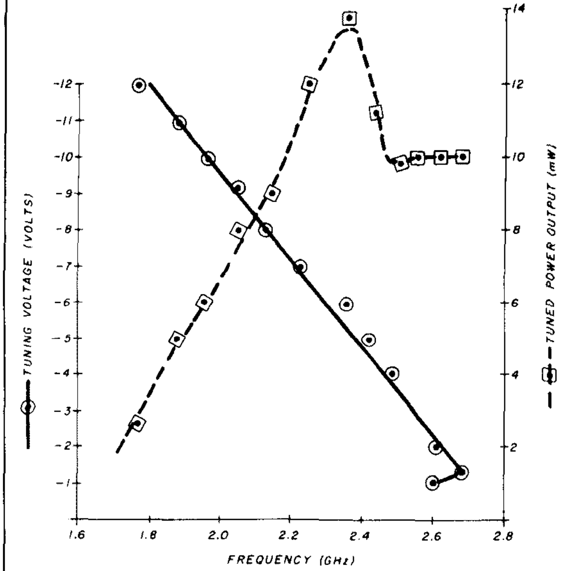

fig. 2. Typical tuning voltage versus frequency and tuned power output versus frequency. Maximum power output occurs when the tuning voltage is between −1 and −7 volts. Tuning is almost a linear function of voltage.

from California Eastern Laboratories, 3005 Democracy Way, Santa Clara, California 95050 or from one of their sales offices.

design specification

Because $V_{COB\ MAX}$ for the NE21935 is 20 volts, 12 VDC power supplies were selected for use with the VCO. The absolute maximum current allowed through the NE21935 is 80 mA. To protect the transistors thermally and provide a wide tuning range, (50 mA, or about 25 mA per transistor,) was selected as the maximum current.

The maximum power dissipation for the NE21935 without case cooling is 390 mW at ambient temperatures up to 50 degrees C (122 degrees F). Since the VCO has an output of 2 to 15 mW with about 50 mW dissipated power, more output is possible at the sacrifice of thermal drift and tuning range. About 25 percent more power output results if the bias voltage is increased to 18 volts. However, tuning range is reduced to about 300 MHz.

Required output power to most mixer diodes is in the range of 5 to 10 mW. Therefore, I wanted at least 10 mW output from the VCO. This power, adequate for communication over a few miles or as excitation power, is easily detectable for test purposes. **Figure 2** shows a typical power output versus frequency plot for the VCO.

I wanted the tuning range of the VCO to be wide enough to cover the entire 13-cm band plus 100 MHz or so on each side, so that a 100 MHz first IF amplifier (FM receiver) could be used in the receive mode. The actual tuning range for VCOs I have tested has been greater than 800 MHz, although when the oscillator is delivering more than 10 mW of power the tuning range decreases to 500 MHz.

I wanted the VCO to have minimum possible phase noise and good stability without generating spurious outputs. To help reduce phase noise the VCO was built on teflon fiberglass PC board and the "hairpin" transmission line was mounted off the PC board with air as the dielectric. The low transistor Q and low hairpin Q would tend to cause high phase noise at this frequency. The VCO should be used in broadband applications only. Spectrum analysis using a homebrew analyzer has shown noise and spurious outputs at least 20 dB below the fundamental.

The shunt capacitance of the tuning voltage circuit was limited to 120 pF to allow for the possibility of modulating the VCO by video signals at a later date.

To prevent stray microwave energy in the shack, it was necessary to enclose the VCO. To help prevent holes or resonances in the tuning range, the enclosure must be less than a quarter wavelength in size in all directions at the highest frequency of oscillation. Because microwave energy can leak through very small cracks in the enclosure, I decided to completely en-

close the VCO in a copper housing, seam soldering all gaps except where the two feedthrough capacitors and the RG-58 coax enter.

The output is closely coupled to the hairpin tank transmission line and therefore high Q and/or mismatched loads at the output may cause tuning holes or resonances. Changes in load will also cause shifts in frequency.

acquiring the parts

The 0.03 inch (0.76 mm) thick, 1 ounce (28 gram), double-sided teflon fiberglass PC board, the eyelets, feedthrough capacitors, and copper foil were purchased from Gateway Electronics, 8123-25 Page Boulevard, St. Louis, Missouri 63130; Surplus Sales, Inc., 2412 Chandler Road, East Bellevue, Nebraska 68005 also stocks needed items. Any value of feedthrough capacitor between 50 and 1000 pF can be used, depending on desired maximum modulation frequency and allowable RF leakage. Standard epoxy fiberglass PC board could probably be substituted for the teflon fiberglass PC board, but some degradation in maximum frequency and some reduction in Q might be expected. All other parts and materials can be obtained from local or mail order dealers.

construction

I use a 20 to 30-watt soldering iron, a 100 to 200-watt soldering iron, a thin track saw (available from model railroad hobby shops), a tubing cutter, rosin core solder, long nosed and Vise-Grip® pliers, a hand drill, a small sharp knife, tin snips, a small heatsink clip, and a low-power magnifying glass for construction. All parts except the PC board, eyelets, the transistor leads, and the copper foil are tinned before soldering. Construction proceeds in this order:

- Prepare the PC board with cut out and eyelets.
- Solder the transistors and "hairpin" to the PC board.
- Install the feedthrough capacitors, resistors, and coax.
- Test the VCO for proper operation.
- Solder the VCO into its enclosure and retest.

The foil pattern shown in fig. 3 is used as the negative to expose a piece of photosensitized 0.03 inch (0.76 mm) thick teflon-fiberglass PC board. The back side of the PC board (the bottom of the enclosure) is covered to prevent etching it away and used as a ground plane. (If you don't have the facilities to etch your own PC board, the etched PC board with eyelets can be obtained from Roensch Microwave, RR #1, Box 156B, Brookfield, Missouri 64628. The price is $5.00, postpaid.)

The PC board is drilled for the eyelets and feedthrough capacitors approximately as shown in **fig. 4.** Locate the holes within 0.1 inch (2.5 mm) of the locations shown so that the entire hole remains at least 0.05 inch (1.3 mm) from the edge of the PC board.

Cut out the space for the hairpin 0.3 inch (8 mm) deep by 0.4 inch (10 mm) wide and then install the two eyelets (or two pieces of 20 gauge wire) for grounding points. Clean the PC board with a non-phosphate cleanser and avoid touching it until all soldering is complete.

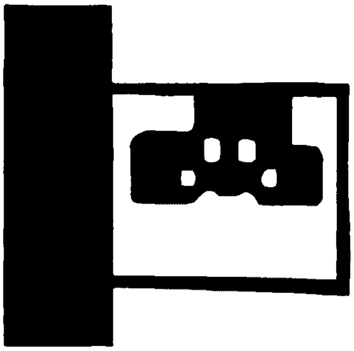

fig. 3. PC board foil pattern (actual size).

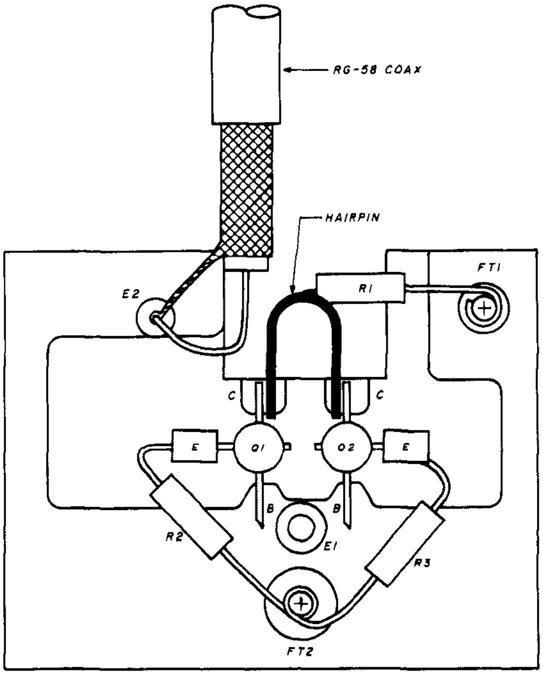

fig. 4. PC board with feedthrough capacitor holes, eyelets installed, and "hairpin" cutout.

Next, install the transistors. (The base lead of the NE21935 is identified by the 45-degree angle of its termination.) The lead opposite the base lead is the collector and the other two leads connect to the emitter. Cut one emitter lead (the opposite emitter lead on each transistor) as short as possible so that the transistors can be mounted close to each other with both base leads pointing in the ground direction as shown in **fig. 5**. With a drop of solder on the tip of a clean, hot 20 to 30 watt soldering iron, place the first transistor on the PC board. While holding a base lead in place with a finger, place a drop of solder on the collector lead and PC board solder pad tacking the lead in place. Make a good solder joint at the other two transistor leads. Then remelt the solder and add a little solder at the collector, base, and emitter to make a good, shiny buildup with adhesion of solder sufficient so no copper shows at all three transistor leads. Do not apply heat at transistor leads for longer than 3 seconds without a long cool-off period. Install the second transistor in the same manner with its cut-off emitter lead about 0.02 inch (0.5 mm) from the other transistor's cut-off emitter lead. The emitter leads must be close but must not touch each other.

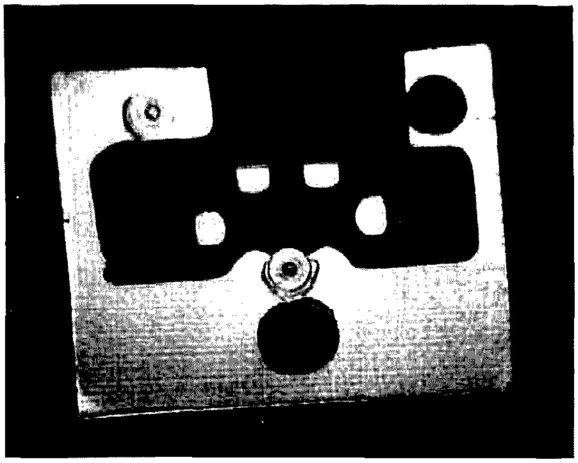

fig. 5. Location of parts on PC board. "Hairpin" is mounted over hairpin cutout. 240-ohm resistor is mounted above hairpin head inside outline of PC board.

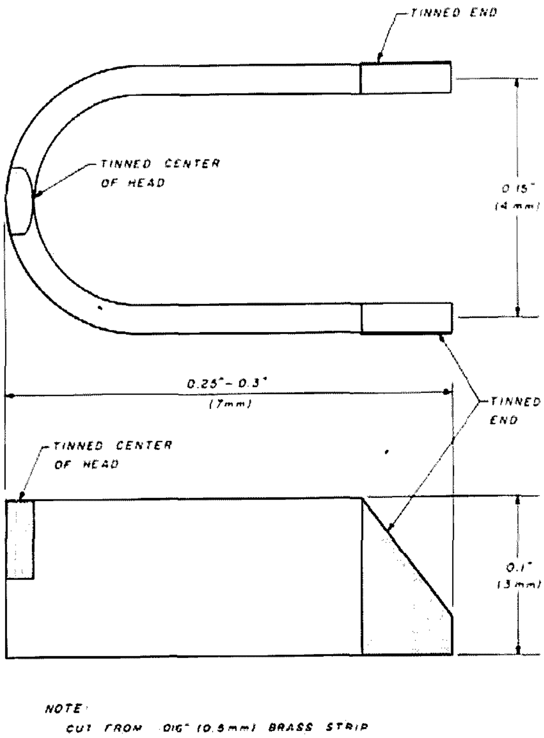

fig. 6. "Hairpin." Areas to be tinned prior to installation are shown.

Cut and bend the hairpin next. I used tin snips to cut a piece of 0.016 inch (0.4 mm) thick brass 0.1 inch (2.5 mm) wide by 0.7 inch (18 mm) long. Bend this piece of brass into its proper shape (**fig. 6**) using long-nosed pliers and your fingers. I make two diagonal cuts at the ends of the hairpin with small diagonal pliers. The sides of the hairpin should be of equal length. Hold the hairpin with Vise-Grip® pliers and tin the outside of each end and the head of the hairpin. A mounted low-power magnifying glass was used for this tinning and for soldering the hairpin to the PC board. Hold the hairpin with a pair of long-nosed pliers in position as shown in **fig. 5**, with the head of the hairpin 0.2 to 0.25 inch (5 to 6 mm) from the PC board. Reheat the solder at one hairpin end and one collector lead securing the hairpin to the PC board. Place a heatsink clip at the head of the hairpin. Then reheat and add solder at the second hairpin end making a solid solder bridge there. Go back to the first hairpin end and add a little solder to make a good solder bridge there also.

The feedthrough capacitors and eyelets are now soldered on both sides of the PC board. Make sure the 0.001 μF feedthrough capacitor is not mounted at an angle; if it is, it could short to the enclosure when it is installed.

Bend one lead of the 240-ohm resistor at a right angle as close to the body of the resistor as possible. Cut this lead to about 0.1 inch (3 mm) leaving the other resistor lead uncut. Use the second resistor lead to hold the resistor while soldering the short lead. (I placed a heatsink clip across the ends of the hairpin and held the resistor at a right angle to and vertically above the hairpin and well inside the outline of the PC board as shown in **fig. 5**.) With the drop of solder on the 20 to 30 watt soldering iron, solder the short resistor lead to the center of the head of the hairpin. The long lead of the resistor should be held against the

0.001 µF feedthrough capacitor while the short lead is soldered. Wrap, cut off, and solder the long lead to the 0.001 µF feedthrough capacitor.

Prepare the two 470-ohm resistors in the same way described for the 240-ohm resistor. Solder the short leads of these two resistors to the emitter leads of the two transistors as shown in **fig. 5**. Wrap, cut off, and solder the long leads of these resistors to the 120 pF feedthrough capacitor.

Prepare a 6-inch (152 mm) piece of RG-58 coax as shown in **fig. 7**, soldering the junction of the shield and center conductor of the coax to the eyelet on the PC board with the pickup loop about 0.02 inch (0.5 mm) from the hairpin as shown in **fig. 5**. The center conductor of the coax must not touch the hairpin. Install a BNC plug on the other end of the coax.

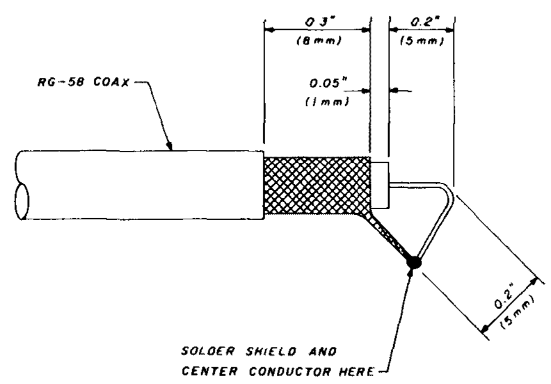

fig. 7. Detail of RG-58 coax pickup loop. Center conductor and shield are soldered together and then soldered to eyelet on PC board.

installing the VCO

Test the VCO as described below. Then cut and bend the enclosure and solder the VCO into it. Cut a piece of 0.02 inch (0.5 mm) flashing copper with tin snips as shown in **fig. 8**. Holding the flat pattern firmly with Vise-Grip® pliers, drill the mounting holes and the hole for the coax. Then cut the copper to make the hole for the coax into a slot.

Cut out a small wooden block 1 × 1.25 × 6 inches (25 × 32 × 156 mm) to use as a jig over which the flat pattern can be bent into a box. Make sure the flat pattern is bent upward because the hole for the RG-58 coax will be on the wrong side if it's bent downward.

Place the tested VCO (PC board) upside down on the bottom of the copper enclosure with the coax in its slot. Tack-solder the PC board in its proper position, placing small pieces of copper foil over the hairpin cutout area and across all gaps between the PC board and the enclosure. Make sure the foil over the hairpin cutout is flat and away from the hairpin. The RG-58 shield should be soldered to the foil and the foil soldered to the enclosure all around. (Do not move the coax until after the enclosure is cool.) The foil can be soldered to the PC board with a 20 to 30 watt iron and to the enclosure with a 100 to 200 watt soldering iron. The completed VCO, ready for final test, is shown in **fig. 9**.

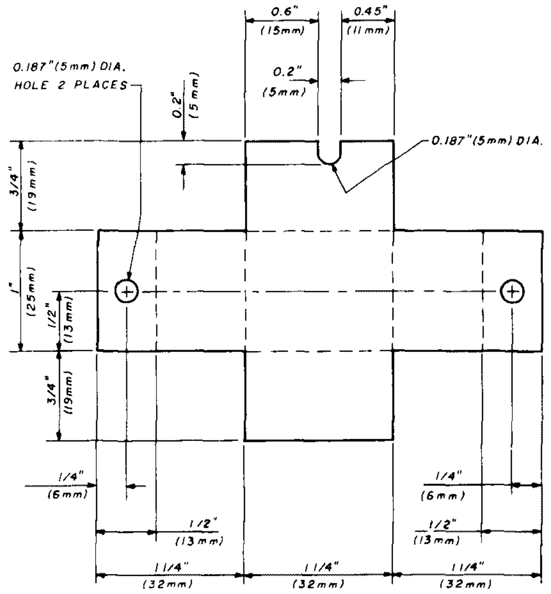

fig. 8. Flat pattern for enclosure uses 2.5 × 3.75 inch (64 × 95 mm) piece of 0.02 inch (0.5 mml thick copper flashing material.

test equipment

A line stretcher is used to determine the approximate frequency; a detector/milliammeter is used as a power indicator. The line stretcher and diode detector can be built as described below.

Use four brass tubes and two teflon-insulated BNC receptacles to make the line stretcher (**fig. 10**). A tubing cutter is used to provide the lengths shown. Place the factory-cut ends of the tubing toward the center of the line stretcher. Slot the larger inner tube and larger outer tube with the track saw at one end in four places (two slots at a time) 90 degrees apart about 1/2 inch (13 mm) deep. Two holes, 5/8 inch (16 mm) and 1/4 inch (6 mm) drilled into a block of wood either nailed down or held in a vise can be used as a jig to hold the tubing while the slots are sawed. This protects the hand holding the tubing in case the thin, sharp track saw slips off the end of the tube. Squeeze the small tubes at their cut ends so that they fit tightly over the center contacts of the BNC receptacles and then solder them in place. Saw the larger

outer tube at the end (with four slots) that you cut with the track saw.

Squeeze the large tubes together uniformly at a slotted (cut) end until the tubes fit tightly over the tinned outer threaded portion of the rear of the BNC receptacles. Then holding the small tube in the center of the larger tube, solder the larger tubes to the threaded portion of the rear of the BNC receptacles.

Gently squeeze the larger inner and outer tubes together so that they fit the smaller inner and outer tube tightly as they are telescoped together. Wipe the outer tube with alcohol to help maintain good electrical contact between the outer tubes.

Build the detector as shown in **fig. 11**. Apply heat for no longer than 5 seconds when soldering the diode. The diode should be tinned at the two solder points before soldering it in place. Keep the leads on the 3-6 pF capacitor and diode as short as possible. The best capacitor to use is one with no leads at all, fabricated from PC board. A piece of 0.03 inch (0.76 mm) thick, double sided teflon/fiberglass PC board 0.3 × 0.3 inch (8 × 8 mm) makes an excellent bypass capacitor for 13-cm microwave frequencies. After the BNC receptacle threads and the PC board are tinned, solder one side of the PC board directly to the BNC receptacle. Heat the receptacle with the 100-watt soldering iron, then quickly solder the bottom side of the PC board to the receptacle with the 20 to 30-watt soldering iron. Then solder the diode in place between the other side of the PC board and the center contact of the BNC receptacle. Commercial 3 to 6 pF capacitors will work but output from the detector will be less. If you desire, the completed assembly can be embedded in silicone caulk to help prevent handling damage, but doing so will reduce output from the detector.

Other types of UHF/microwave diodes may be used if their junction capacitance is less than 1 pF. The HP5082-2835 diode is available from Radio Shack (Part No. 276-1124). 1N21 or 1N23 diodes are available from MHz Electronics, 2111 W. Camelback Road, Phoenix, Arizona 85015.

testing

The overall test setup is shown in **fig. 12**. I keep the RG-58 cables less than 6 inches (150 mm) long because RG-58 is very lossy at 13 cm. With the tuning voltage adjusted to a preliminary setting of −5 VDC, adjust the line stretcher for a minimum reading on the milliammeter. Mark this position on the smaller outer tube of the line stretcher with a felt tip pen. The line stretcher is then adjusted longer to the next minimum as indicated on the milliammeter and the smaller

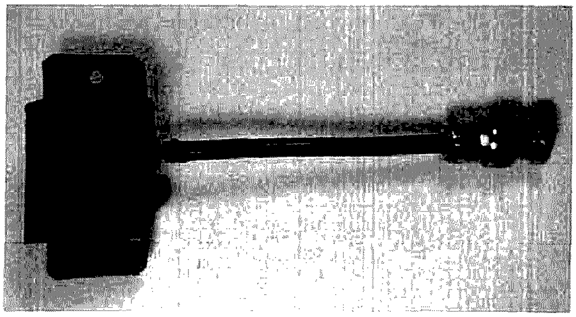

fig. 9. The completed microwave VCO.

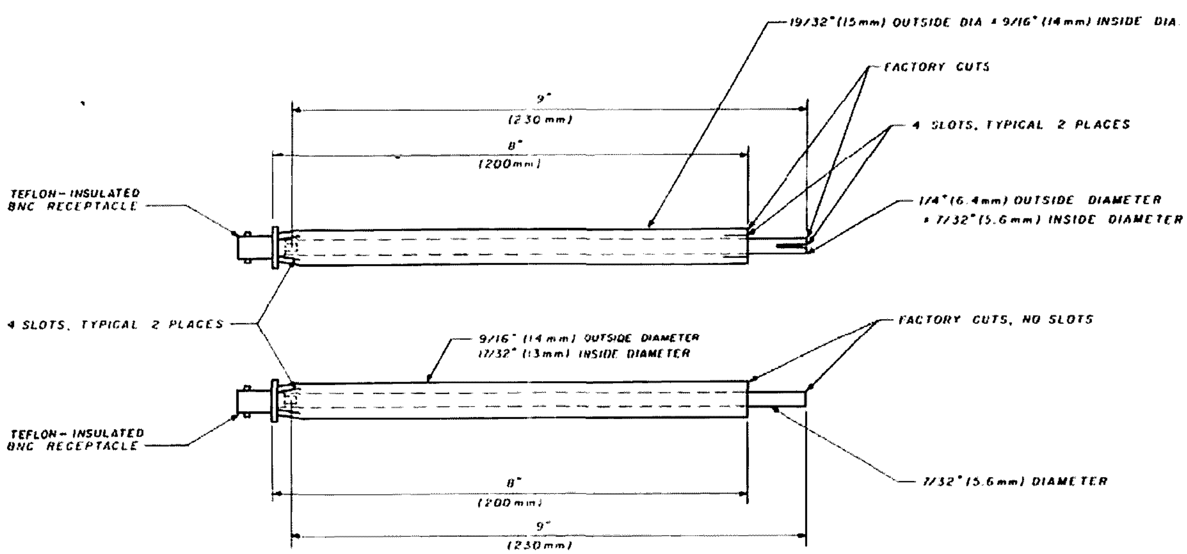

fig. 10. The line stretcher is built from four 0.016 inch (0.4 mm) wall brass tubes and two BNC receptacles.

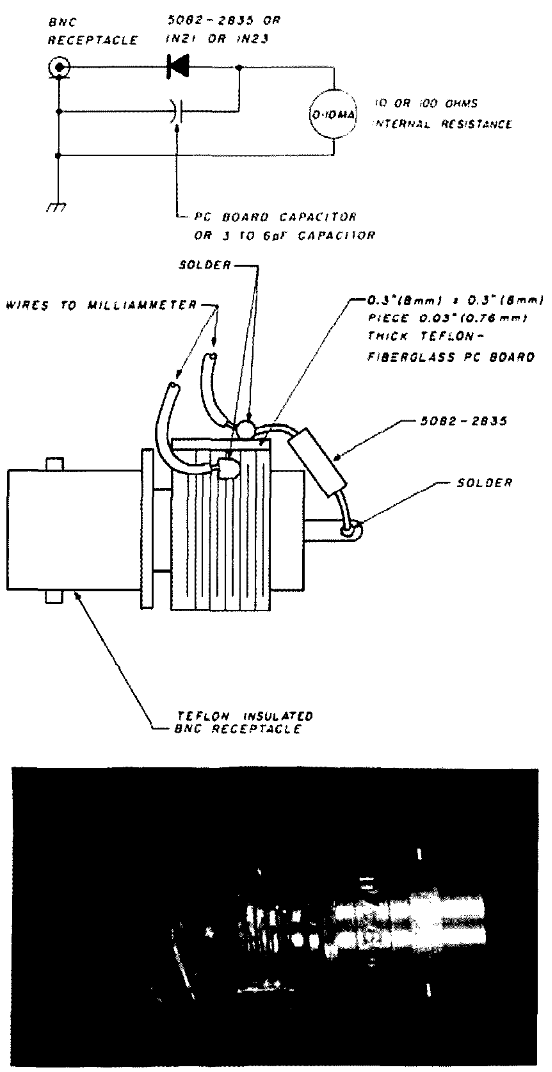

fig. 11. The diode detector is built on a teflon insulated BNC receptacle using a microwave diode and homemade bypass capacitor.

outer tube marked again. Some shift in frequency of the VCO (10 or 20 MHz) occurs as the line stretcher is lengthened, but the frequency is approximately the same at each minimum. The distance (*d*) in cm between these two marks is equal to the wavelength divided by 2. The frequency (*f* in MHz) is equal to 15,000 divided by *d*. With careful adjustment and measurement, it should be possible to tune the oscillator to the center of the upper 13-cm band (12.4 cm). Since the total tuning range of this VCO is about 1.8 to 2.6 GHz, voltages between about −4 and −6 volts should tune the entire 13-cm band.

Frequency can be determined more accurately by injecting a known frequency (or harmonic of a known

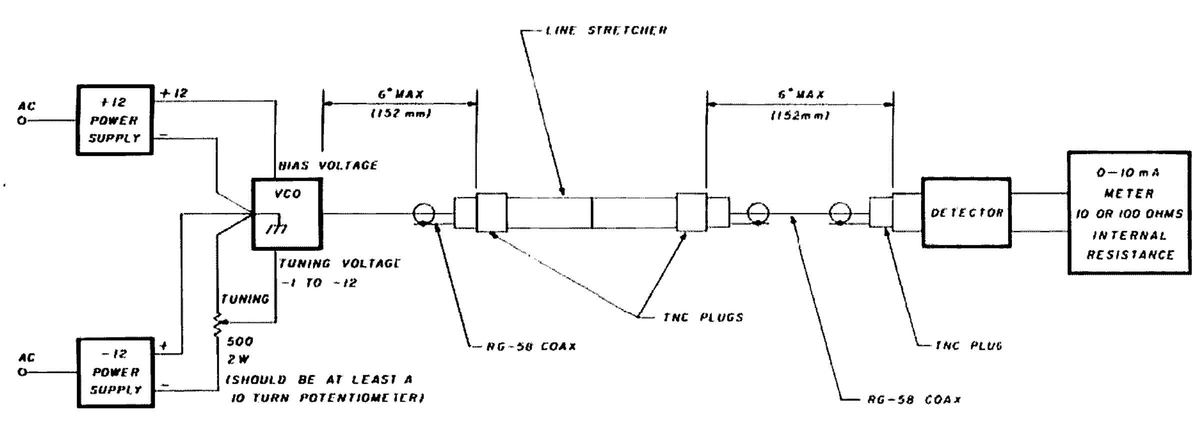

fig. 12. Test configuration.

frequency) into the detector area with a probe and tuning the VCO until a zero beat indication is displayed on an oscilloscope. Frequency modulation of the VCO or frequency standard will cause a "birdie" to appear on the oscilloscope if the horizontal sweep of the oscilloscope is synchronized with the modulation frequency. This makes it easier to find the VCO frequency since tuning is very rapid through the 2 or 3 MHz bandwidth of the oscilloscope.

To determine approximate power output, adjust the line stretcher for maximum current through the detector. An approximate level of output power can be determined by using **table 1**. (For more information on estimation of power output see reference 1.)

The line stretcher used above can be used to help match the output of this VCO to any load that might be used. Although this matching is not as good as can be realized with a double stub tuner, it may be adequate in many cases.[2] Matching can be improved — or made worse — by adjusting the position of the coupling loop in relation to the "hairpin" and/or adjusting the size of the coupling loop. The length or position of the "hairpin" on the solder pads may be adjusted to increase the output over a particular band of frequencies. Maximum output occurs with a tuning voltage between about −1 and −7 volts. Bias voltage can also be adjusted to change frequency and output power.

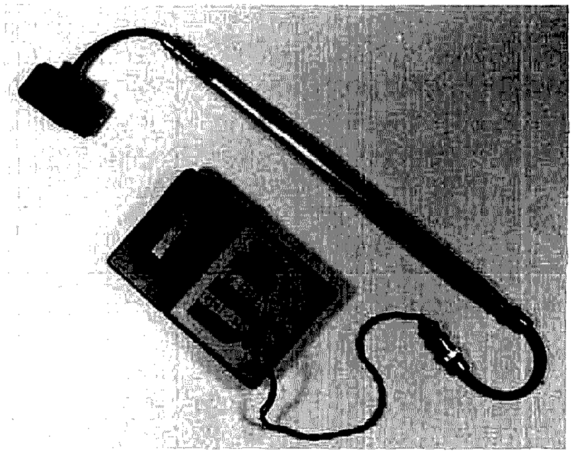

table 1. Correlation of VCO output power with current meter reading.

meter reading with Schottky diode (such as 5082-2835) (mA)	meter reading with contact diode (such as 1N21 or 1N23) (mA)	approximate power out (mW)
10-20	9-10	10-20
6-10	4-6	5-10
0.5-0.8	0.8-0.9	1

safety

There is little danger from the output of this microwave VCO since maximum power output is about 15 mW. (OSHA sets 10 mW/cm² as the maximum safe radiation density.) If the maximum 15 mW output of this VCO were concentrated in an area of less than 1.5 cm² there could be a hazard. This might happen with a waveguide horn, parabolic antenna, or other type of high gain device. When working with microwaves never look into an active waveguide, antenna, or other high gain device. Never place your head at the focal point of an irradiated high gain antenna. Microwave radiation above 10 mW/cm² can harm your eyes. *Never expose your body to high-level radiation*.

applications

This VCO has many uses. It can be used as a local oscillator, as a low-power FM transmitter or exciter,

as a component in a spectrum analyzer or sweep generator, as a signal source to test amplifiers, attenuators, antennas, transistors, filters, oscillators, detectors, feed horns, or mixers. It can also be used at slightly lower frequencies for TVRO reception if a doubling mixer is used.[3]

This VCO cannot, however, be used with narrowband systems where crystal stability and minimum phase noise are required. If accurate frequency control is needed, a phase-locked loop and frequency counter may be added to the VCO.[4]

If you have problems with the construction, testing, or operation of this VCO I'll be glad to help you. Please send a SASE with your inquiry to me at the address shown at the beginning of this article.

references

1. Evans & Jessop, *VHF-UHF Manual*, (RSGB), 1982, page 8.16.
2. George Hatherell, "Double-stub Tuner for 1296 MHz," *ham radio*, December, 1978, page 72.
3. Jim Dietrich, "Twin-diode Mixer — A New Microwave Mixer," *ham radio*, October, 1978, page 84.
4. Norman J. Foot, "Multipurpose Voltage-tuned UHF Oscillator," *ham radio*, December, 1980, page 12.

ham radio

✓ 108

trade off power
for antenna gain at VHF?

These computer programs simplify calculations and help you decide

If you operate at VHF or UHF frequencies and would like to find the most cost-effective way to increase your range — or if you have $100 to spend but don't know whether to invest it in higher power, more antenna gain, or a taller tower — a few minutes at your computer can give you projected communications ranges for any combination you wish to try. While programs shown in **fig. 1** and 2 are written for Commodore-64 and Texas Instruments-99/4A computers, respectively, they can be easily converted, using BASIC, for use on other microcomputers. Both free space range (for satellite communications) and range over real earth are given.

propagation curves

Although propagation curves have been available for about 40 years, most hams have either not been aware of them or have not known how to use them. The classical curves developed by Bell Labs cover frequencies from 200 kHz to 600 MHz, distances of 0.5 to 1000 miles, and are arranged in six sections covering propagation over sea water, good soil, and poor soil for vertical and horizontal polarizations.[1] Typical inputs and outputs are expressed in terms of 1 kW transmitted from a grounded vertical antenna and units of field strength in dB above 1 microvolt per meter, however, and one must be wise in the ways of antenna conversions to use them. Predictions are also complicated at HF, where antennas are usually located within a few wavelengths of ground, because actual antenna directivity and efficiency are directly affected by soil conductivity.

Propagation predictions at VHF and UHF are more straightforward because antennas at these frequencies are usually mounted many wavelengths above ground. At these frequencies the communications range is essentially independent of polarization used and soil type for antenna heights of 100 feet or more, and variations of not more than 3 or 4 dB are to be expected at heights of 25 feet. Based upon these facts (and other assumptions), the ESSA curves are useful for frequencies above 100 MHz.[2]

The computer programs described below utilize data taken from selected curves in reference 2. After you enter operating frequency, receiver sensitivity, transmitter power, and antenna heights and gains, the programs calculate your expected communications range. A typical output from the Commodore-64 program is shown in **fig. 1A**.

C-64 program description

This section describes the program as written for the Commodore-64 because the machine is very popular among hams and because the program contains the most features. (Similar versions for other computers, with fewer features, will be described later.) In the Commodore-64 program shown in **fig. 1, line**

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fig. 1. Communications range calculation program for the Commodore 64.

```
2 GOSUB 500
10 PRINTCL$;PRINT"        VHF/UHF PROPAGATION PROGRAM
12 PRINT:PRINT"           FOR THE COMMODORE 64
14 PRINT:PRINT"               VI.0 C 1984 BY
16 PRINT:PRINT"                LYNN A. GERIG
20 FORJ=1TO10:PRINT:NEXT
22 PRINT"TO CHANGE BORDER, SCREEN, OR LETTER
24 PRINT"COLORS, PRESS B, S, OR L, RESPECTIVELY.
26 PRINT:PRINT"TO EXIT TO PROGRAM, PRESS <RETURN>.
30 GETA$:IFA$=""THEN30
32 IFA$="B"THENPOKE53280,(PEEK(53280)AND15)+1
34 IFA$="S"THENPOKE53281,(PEEK(53281)AND15)+1
36 IFA$="L"THENPOKE646,(PEEK(646)AND15)+1:GOTO10
38 IFA$=CHR$(13)THEN50
40 GOTO30
50 PRINTCL$"THIS PROGRAM WILL CALCULATE EXPECTED
52 PRINT"RANGES FOR VHF (100-175 MHZ) AND UHF
54 PRINT"(225-500 MHZ) FREQUENCIES.  APPROXIMATE
56 PRINT"DYNAMIC RANGE IS FOR PATH LOSSES OF 125
58 PRINT"TO 200 DB, COVERING MOST APPLICATIONS
60 PRINT"FOR RCVR SENS .5 TO 10 MICRO-VOLTS AND
62 PRINT"XMTR POWER OF 1 TO 1000 WATTS.  PROGRAM
64 PRINT"COVERS ANTENNA HEIGHTS FROM 25 FT TO     100,000 FT.
70 PRINT:PRINT"PROGRAM DEFAULTS TO RCVR SENS AND XMTR
72 PRINT"PWR IN DBM.  WOULD YOU RATHER WORK WITH
74 PRINT"MICRO-VOLTS AND WATTS (Y=YES)";:INPUT D$
100 PRINTCL$"PRESS <V> FOR VHF OR <U> FOR UHF":PRINT
102 GETF$
104 IFF$="V"THENPRINT"ENTERING VHF DATA":GOSUB2000:GOTO200
106 IFF$="U"THENPRINT"ENTERING UHF DATA":GOSUB3000:GOTO200
108 GOTO102
199 :
200 PRINT:GOSUB600:REM SELECT FREQUENCY
205 PRINT:GOSUB700:REM SELECT XMTR PWR & RCVR SENS
210 PRINT:GOSUB900:REM SELECT ANTENNA GAINS
215 PRINT:GOSUB800:REM SELECT ANTENNA HEIGHTS
220 :
300 PRINTCL$F$"HF PROPAGATION:  FREQ ="F"MHZ
302 PRINT:PRINT"TRANSMITTER POWER OUT:"PD;TAB(30);"DBM
304 PRINTTAB(22)PW;TAB(30)"WATTS"
306 PRINT:PRINT"RECEIVER SENSITIVITY: "RD;TAB(30)"DBM
308 PRINTTAB(22)RM;TAB(30)"UV
310 PRINT:PRINT"LOWER ANTENNA : "GL"DBI @"H1"FT
312 PRINT"UPPER ANTENNA : "GU"DBI @"H2"FT
314 PRINT:PRINT"COAXIAL LINE LOSSES : "LL"DB
320 PL=PD-RD+GU+GL-LL
322 DF=PL-37-20*LOG(F)/LOG(10)
324 DF=10^(DF/20):DF=INT(DF+.5)
326 PRINT:PRINTPL"DB PATH
328 PRINT"FREE SPACE PATH ="DF"MILES
350 PL=PL-20*LOG(F/F1)/LOG(10):PRINT
352 IFPL<P1THENPRINT"RANGE NOT IN PROGRAM:  <"S1"MILES
354 IFPL<P1THENPRINTTAB(25)"<"D1"(NAUT MI)":GOTO400
356 IFPL>P5THENPRINT"RANGE NOT IN PROGRAM:  >"S5"MILES
358 IFPL>P5THENPRINTTAB(23)">"D5"(NAUT MI)":GOTO400
360 IFPL>=P1ANDPL<P2THEN DN=D1+(D2-D1)*(PL-P1)/(P2-P1)
362 IFPL>=P2ANDPL<P3THEN DN=D2+(D3-D2)*(PL-P2)/(P3-P2)
364 IFPL>=P3ANDPL<P4THEN DN=D3+(D4-D3)*(PL-P3)/(P4-P3)
366 IFPL>=P4ANDPL<P5THEN DN=D4+(D5-D4)*(PL-P4)/(P5-P4)
368 DS=INT(DN*1.151+.5):DN=INT(DN+.5)
370 PRINT"MAXIMUM EXPECTED RANGE: "DS"MILES
372 PRINTTAB(24)DN"(NAUT MI)
400 PRINT:PRINT:PRINTLL$
402 PRINT"G=MODIFY ANT GAINS         R=RUN AGAIN
404 PRINT"H=MODIFY ANT HEIGHTS       P=PRINTER DUMP
406 PRINT"X=MODIFY R/T SENS/PWR      Q=QUIT
408 PRINT"F=NEW FREQ (SAME BAND)";
410 FORJ=1TO10:GETA$:NEXT
412 GETA$:IFA$=""THEN412
413 IFA$="P"THEN430
414 PRINT:PRINT
415 IFA$="R"THEN100
416 IFA$="Q"THENSYS65126
418 IFA$="G"THENGOSUB900:GOTO300
420 IFA$="H"THENGOSUB800:GOTO300
422 IFA$="X"THENGOSUB700:GOTO300
424 IFA$="F"THENGOSUB600:GOTO300
426 GOTO412
430 REM SCREEN DUMP TO PRINTER
432 OPEN3,3:OPEN4,4:PRINTHO$;:PRINT#4:PRINT#4,LL$
434 FORJ=0TO759:GET#3,A$:PRINT#4,A$;:NEXT:PRINT#4,LL$
436 CLOSE4:CLOSE3:FORJ=1TO5:PRINT:NEXT:GOTO412
500 HO$=CHR$(19):CL$=CHR$(147):D$="N":DIMH$(13,15):DIMH(15)
502 H(1)=25:H(2)=50:H(3)=100:H(4)=500:H(5)=1000:H(6)=2000
504 H(7)=5000:H(8)=10000:H(9)=15000:H(10)=20000:H(11)=30000
506 H(12)=40000:H(13)=60000:H(14)=80000:H(15)=100000
508 LL$=" ---------------------------------------"
510 RETURN
600 PRINT"FREQUENCY IN MHZ ("FL"-"FU")";:INPUTF
602 IFF<FLORF>FUTHEN600
604 RETURN
700 IFD$="Y"THENINPUT"INPUT XMTR POWER (IN WATTS)";PW:GOTO710
702 INPUT"INPUT XMTR POWER (IN DBM)";PD
704 PW=(PD-30)/10:PW=10^PW
706 IFPW>=1THENPW=INT(PW*10+.5)/10:GOTO720
708 IFPW<1THENPW=INT(PW*1000+.5)/1000:GOTO 720
710 PD=10*LOG(PW)/LOG(10)+30:PD=INT(PD*10+.5)/10
720 IFD$="Y"THENINPUT"RCVR SENSITIVITY (IN MICRO-VOLTS)";RM:GOTO730
722 INPUT"RCVR SENSITIVITY (IN DBM)";RD
723 IFRD>0THENPRINT"<1 MW IS A NEGATIVE NUMBER":GOTO722
724 RM=(RD+107)/20:RM=10^RM
726 IFRM>=1THEN RM=INT(RM*10+.5)/10:GOTO740
728 IFRM<1THEN RM=INT(RM*100+.5)/100:GOTO740
730 RD=20*LOG(RM)/LOG(10)-107:RD=INT(RD*10+.5)/10
740 RETURN
800 REM ANTENNA HEIGHTS
802 PRINTCL$"CHOOSE ANTENNA HEIGHTS BY SELECTING
804 PRINT"NUMBERS FROM THE FOLLOWING MENU":PRINT:PRINT
806 PRINT"1 =   25'    6 =   2000'    11 =  30000'
808 PRINT"2 =   50'    7 =   5000'    12 =  40000'
810 PRINT"3 =  100'    8 =  10000'    13 =  60000'
812 PRINT"4 =  500'    9 =  15000'    14 =  80000'
814 PRINT"5 = 1000'   10 =  20000'    15 = 100000'
820 PRINT:PRINT
830 INPUT"SELECT HEIGHT OF LOWER ANTENNA";H1
832 INPUT"SELECT HEIGHT OF UPPER ANTENNA";H2
834 IFH1<1ORH2<1ORH1>15ORH2>15THENPRINT"NOT IN MENU":GOTO830
836 IFH1>H2THENPRINT"LOWER AND UPPER REVERSED":GOTO830
840 H$=H$(H1,H2):H1=H(H1):H2=H(H2)
850 P1=VAL(MID$(H$,1,3)):D1=VAL(MID$(H$,4,3))
852 P2=VAL(MID$(H$,7,3)):D2=VAL(MID$(H$,10,3))
854 P3=VAL(MID$(H$,13,3)):D3=VAL(MID$(H$,16,3))
856 P4=VAL(MID$(H$,19,3)):D4=VAL(MID$(H$,22,3))
858 P5=VAL(MID$(H$,25,3)):D5=VAL(MID$(H$,28,3))
860 S1=INT(D1*1.151+.5)
862 S5=INT(D5*1.151+.5)
870 PRINTCL$:RETURN
900 INPUT"GAIN OF LOWER ANTENNA (IN DBI)";GL
902 INPUT"GAIN OF UPPER ANTENNA (IN DBI)";GU
904 INPUT"COAXIAL LINE LOSSES (IN DB)";LL
906 RETURN
2000 F1=125:FL=100:FU=175:REM VHF DATA
2005 H$( 1, 1)="140010160025170038176080228340
2010 H$( 1, 2)="133010160030170046175068228560
2015 H$( 1, 3)="130010160037170085175075225388
2020 H$( 1, 4)="117010162060170075175100225408
2025 H$( 1, 5)="113010165070169088173100228418
2030 H$( 1, 6)="103010158080170100175120225425
2035 H$( 1, 7)="100010140085170130177160218400
2040 H$( 1, 8)="118060132100170168176180213400
2045 H$( 1, 9)="125100150160175195185260210400
2050 H$( 1,10)="126120145170165200175215215450
2055 H$( 1,11)="130160150210177250186320210440
2060 H$( 1,12)="128180150240177288188360210470
2065 H$( 1,13)="128220145280178330188400210810
2070 H$( 1,14)="132270150330177368189440210580
2075 H$( 1,15)="132500150565178410189480210880
2080 H$( 2, 2)="128010158030168048170060210300
2085 H$( 2, 3)="120010150032168080175108210310
2090 H$( 2, 4)="110010158060167080183180208300
2095 H$( 2, 5)="108010160075168100180170207320
2100 H$( 2, 6)="110030163090170120183200208340
2105 H$( 2, 7)="115080135088162128168140210370
2110 H$( 2, 8)="120090148140165160170175210400
2115 H$( 2, 9)="125120150165170195180260210420
2120 H$( 2,10)="125140140170171215188300210440
2125 H$( 2,11)="130180145215171250182320208440
2130 H$( 2,12)="150210150250173285185360211500
2135 H$( 2,13)="130255150300174335187420206810
2140 H$( 2,14)="130290180340173370190470208560
2145 H$( 2,18)="135340150375174410185480208600
2150 H$( 3, 3)="125015155045166065174120203280
2155 H$( 3, 4)="113020152060163080177160203300
2160 H$( 3, 5)="110020152070165100195260210355
2165 H$( 3, 6)="110030130000164110195275209360
2170 H$( 3, 7)="115060140100164135196300211400
2175 H$( 3, 5)="125110158160167175184270204380
2180 H$( 3, 9)="125130140160165195172220205410
2185 H$( 3,10)="125150135170165215172240205430
2190 H9( 3,11)="125188148220168250173280210800
2195 H$( 3,12)="130220138240165280175320203480
2200 H$( 3,13)="130270145300160320170340203520
2205 H$( 3,14)="135320165370175400180440203560
2210 H$( 3,15)="133355163400175435178460204600
2215 H$( 4, 4)="110050150080159095165120204330
2220 H$( 4, 5)="115045151090160110170158204340
2225 H$( 4, 6)="116060155110161125176200204350
2230 H$( 4, 7)="123100156145165165190295210420
2235 H$( 4, 8)="125135157180165200195350210450
2240 H$( 4, 9)="130170160210168235199400218500
2245 H$( 4,10)="130190158225168260195400210495
2250 H$( 4,11)="130225160260170295184375205485
2255 H$( 4,12)="135260160294169220183400203500
2260 H9( 4,13)="125180129300163350172380202540
2265 H$( 4,14)="125180130340162385173420205600
2270 H$( 4,15)="125180132380165430174460204620
2275 H$( 5, 3)="116065144090155110165140203340
2280 H$( 5, 6)="120080145110160135170180204360
2285 H$( 5, 7)="125120145140158160165180203380
2290 H$( 5, 8)="125150150180162200167215202400
2295 H$( 5, 9)="125174150205160220165234201420
2300 H$( 5,10)="127200150225159240166260200435
2303 H$( 5,11)="129235155265160275165286202480
2310 H$( 5,12)="125180128266160505167325201800
2315 H$( 5,13)="125180130315157250168374200540
2320 H$( 5,14)="128260131355150380168410203600
2325 H$( 5,15)="128260132395160430170438199600
2330 H$( 6, 6)="125100130130160150168180200350
2335 H$( 6, 7)="122130149160160180170220200375
2340 H$( 6, 8)="128166154200160210168240200405
2345 H$( 6, 9)="125190150220160233168265200430
2350 H$( 6,10)="126212150240160260167275200445
2355 H$( 6,11)="129250153280163300170325200480
2360 H$( 6,12)="125180129285150306166335200510
2365 H$( 6,13)="125180130330155360168390208590
2370 H$( 6,14)="127220131370135403169430205620
2375 H$( 6,15)="127220132410155440168465200625
2380 H$( 7, 7)="125165150195160210169240200400
2385 H$( 7, 8)="125200143220157240166260200430
2390 H$( 7, 9)="127225155260164280170300200460
```

```
2395 H8( 7,10)="128250153280164300170320201480
2400 H8( 7,11)="130285150310160325166340200510
2405 H8( 7,12)="125180129315150340167370200535
2410 H8( 7,13)="127230130365156395170425200580
2415 H8( 7,14)="127230132405158440170465200620
2420 H8( 7,15)="127230133445155470170500200650
2425 H8( 8, 8)="129235155265163280168300200460
2430 H8( 8, 9)="130260155290160300168320203500
2435 H8( 8,10)="130280155310164330169350200505
2440 H8( 8,11)="130320150340162360170385200540
2445 H8( 8,12)="125180131350157380170410200565
2450 H8( 8,13)="125180132395160435170460200605
2455 H8( 8,14)="127230133440156470170500200650
2460 H8( 8,15)="127230133475160510175555205720
2465 H8( 9, 9)="130290145305157320166340200505
2470 H8( 9,10)="130305155340165360170375205360
2475 H8( 9,11)="130345156375167400175430205595
2480 H8( 9,12)="125180131374155405170440200590
2485 H8( 9,13)="127240132425159460170485201640
2490 H8( 9,14)="130320133465160505170525200675
2495 H8( 9,15)="130320133500160540175580200710
2500 H8(10,10)="125180130330165380175415202560
2505 H8(10,11)="125180132370155395170430205613
2510 H8(10,12)="127230132395160435175475205640
2515 H8(10,15)="129280133450155475172510201660
2520 H8(10,14)="130300133490155515172550205725
2525 H8(10,15)="130300134525165570176620205760
2530 H8(11,11)="125180133400160440175480205650
2535 H8(11,12)="130300133435155460172500205675
2540 H8(11,13)="130500134485165530180580205725
2545 H8(11,14)="130300135525160560175600205760
2550 H8(11,15)="130300135560165805180655205800
2555 H8(12,12)="130300133480163505175540205700
2560 H8(12,13)="130300134515167560177600205750
2565 H8(12,14)="130300138855165600178640210820
2570 H8(12,15)="130300135590160625175660205825
2575 H8(13,13)="130300135560162600175635200760
2580 H8(13,14)="130300135605160635175675200805
2585 H8(13,15)="130300136640161675177715200840
2590 H8(14,14)="130300136645170700185760208900
2595 H8(14,15)="130300137685162720177760210940
2600 H8(15,15)="130300137720163760179800210980
2610 RETURN
3000 F1=300:FL=225:FU=500:REM UHF DATA
3005 H8( 1, 1)="142010170030178050182100225315
3010 H8( 1, 2)="136010150020170035178060220300
3015 H8( 1, 3)="130010150025170045176080215275
3020 H8( 1, 4)="125020188055178070188125210280
3025 H8( 1, 5)="120020165065175080185135210270
3030 H8( 1, 6)="127040145060170085177100215305
3035 H8( 1, 7)="127060140080174120178130210305
3040 H8( 1, 8)="126085140110176155180166200280
3045 H8( 1, 9)="132120145145175180181190200300
3050 H8( 1,10)="135145150170175195180210215400
3055 H8( 1,11)="135175145200177235181245200355
3060 H8( 1,12)="140220150235175265183280200380
3065 H8( 1,13)="140260150285180320185330200430
3070 H8( 1,14)="140300150325179360185370200465
3075 H8( 1,15)="140333151360180395187410200500
3080 H8( 2, 2)="132010155030168040175060215280
3085 H8( 2, 3)="125010150030164045175070213280
3090 H8( 2, 4)="120020165060170070175090210270
3095 H8( 2, 5)="120030165070171080175096210280
3100 H8( 2, 6)="120040137060167090173100210290
3105 H8( 2, 7)="125070140090170125176135210320
3110 H8( 2, 8)="131110140125170155176175205320
3115 H8( 2, 9)="135135151160170180177190205340
3120 H8( 2,10)="135160150180170205174220205360
3125 H8( 2,11)="140200145210175240180255205400
3130 H8( 2,12)="140230147240172270180285205425
3135 H8( 2,13)="140275148290173320181335200445
3140 H8( 2,14)="137308148330175365184380200480
3145 H8( 2,15)="139350150370175395184420197500
3150 H8( 3, 3)="120010140025160045170060205240
3155 H8( 3, 4)="118020160060170075183140210280
3160 H8( 3, 5)="120030160070167080173100205260
3165 H8( 3, 6)="122050155080168100175120210500
3170 H8( 3, 7)="130080155110170130175145210525
3175 H8( 3, 8)="135120168160173170177180210360
3180 H8( 3, 9)="132140166180175195180220210380
3185 H8( 3,10)="132160145180165200177220210400
3190 H8( 3,11)="135200150220170245177255215460
3195 H8( 3,12)="133230150250173280180300210460
3200 H8( 3,13)="138280150300172320180345210500
3205 H8( 3,14)="139320150340170360180580210540
3210 H8( 3,15)="140355155380172400180415210580
3215 H8( 4, 4)="120040155070165085175115205265
3220 H8( 4, 5)="125055153080167100175125210305
3225 H8( 4, 6)="125072157100170120180160210315
3230 H8( 4, 7)="128105160135170150178180210345
3235 H8( 4, 8)="132145155165168180177200215400
3240 H8( 4, 9)="135170158190172210180240210395
3245 H8( 4,10)="135192145215175240180260210420
3250 H8( 4,11)="135225145255175275180290210450
3255 H8( 4,12)="138260160280172295180320215500
3260 H8( 4,13)="140355155325170340180365210520
3265 H8( 4,14)="140352160370171385180206210560
3270 H8( 4,15)="140390154400170420180240210590
3275 H8( 5, 5)="124065160100171120177140210310
3280 H8( 5, 6)="127085160115170130178160206300
3285 H8( 5, 7)="130115160144170160177180210350
3290 H8( 5, 8)="132150155170168190176206210380
3295 H8( 5, 9)="133175159200170215175230210400
3300 H8( 5,10)="133200158220170235178260210425
3305 H8( 5,11)="135232161260173280179300210455
3310 H8( 5,12)="136265160290171305180330210485
3315 H8( 5,15)="139520160340174360180380210530
3320 H8( 5,14)="140360160380174400180415210560
3325 H8( 5,15)="140394165420175436182455210600
3330 H8( 6, 6)="128105149120166140176166210340
3335 H8( 6, 7)="132135159160170175177195210360
3340 H8( 6, 8)="134170156190171210178230211400
3345 H8( 6, 9)="134195161220170232177250211420
3350 H8( 6,10)="136220160240171255179280211440
3355 H8( 6,11)="138254165280173295179315210470
3360 H8( 6,12)="137285155300170315178340211500
3365 H8( 6,13)="138330165360174375181395210540
3370 H8( 6,14)="140375164400175415181430210580
3375 H8( 6,15)="143415167440176455184480210610
3380 H8( 7, 7)="132165165200175220180240205363
3385 H8( 7, 8)="135205156220170240180270210420
3390 H8( 7, 9)="136230160250170265180295210445
3395 H8( 7,10)="130130135245165275176300210460
3400 H8( 7,11)="131140137180168315178335210495
3405 H8( 7,12)="133200138315165345180375210525
3410 H8( 7,13)="133200139368170400182430212580
3415 H8( 7,14)="135240140410170440180465210610
3420 H8( 7,15)="135240140445165470180495210640
3425 H8( 8, 8)="130140135235165260180300210450
3430 H8( 8, 9)="130140136255165290177310210470
3435 H8( 8,10)="130140137280165310178340205470
3440 H8( 8,11)="130140138320165346178370210520
3445 H8( 8,12)="135240138350168380179405210550
3450 H8( 8,13)="135240139400170430180455210600
3455 H8( 8,14)="135240140445170470180495210640
3460 H8( 8,15)="137300141480170510182540210670
3465 H8( 9, 9)="132160137286169320183360210500
3470 H8( 9,10)="132160138310169340183380210515
3475 H8( 9,11)="135240138347170380180405210550
3480 H8( 9,12)="135240139375163400182440210575
3485 H8( 9,13)="135240140425170460181485210620
3490 H8( 9,14)="135240141465170500181525210660
3495 H8( 9,15)="138340142505165530181560210695
3500 H8(10,10)="135240138330167360180390210530
3505 H8(10,11)="135240138370168400160425208560
3510 H8(10,12)="135240139395170430182460210595
3515 H8(10,13)="135240140446170480185520210640
3520 H8(10,14)="137300141490170520180545210680
3525 H8(10,15)="137300142525170555185595210715
3530 H8(11,11)="135240140405165430180455210600
3535 H8(11,12)="135240140430165460181490210630
3540 H8(11,13)="137300141505172520183550210670
3545 H8(11,14)="137300142528172560185590210715
3550 H8(11,15)="138300143565170592185630210750
3555 H8(12,12)="140400141465175505184530210640
3560 H8(12,13)="140400142515172550187590210705
3565 H8(12,14)="140400143555165580185620209740
3570 H8(12,15)="140400143590175635188670210780
3575 H8(13,13)="138360142565170595186650210755
3580 H8(13,14)="138360143608170640185670210790
3585 H8(13,15)="138360144645178690191730210830
3590 H8(14,14)="140420143647170680185710210830
3595 H8(14,15)="140420144686170715188755210870
3600 H8(15,15)="140400145720170750190800210900
3610 RETURN
```

```
--------------------------------------------
VHF PROPAGATION:  FREQ = 146.5 MHZ

TRANSMITTER POWER OUT: 44.8    DBM
                       30      WATTS

RECEIVER SENSITIVITY: -110.7   DBM
                       .65     UV

LOWER ANTENNA :   12.4 DBI @ 50 FT
UPPER ANTENNA :   7.5 DBI @ 100 FT

COAXIAL LINE LOSSES :   1.6 DB

 173.8 DB PATH
FREE SPACE PATH = 47224 MILES

MAXIMUM EXPECTED RANGE:   105 MILES
                          91 (NAUT MI)

--------------------------------------------
G=MODIFY ANT GAINS          R=RUN AGAIN
H=MODIFY ANT HEIGHTS        P=PRINTER DUMP
X=MODIFY R/T SENS/PWR       Q=QUIT
F=NEW FREQ (SAME BAND)
--------------------------------------------
```

fig. 1A. Typical output from Commodore-64 program.

2 sends you to a subroutine (lines 500-510) where arrays are dimensioned and certain variables are established. The actual program begins with line 10.

Lines 10 through **40** display the program title on the screen and provide the opportunity to select any combination of screen, border, and letter colors — assuming, of course, that you're using a color monitor — that you find pleasing. You're not stuck with the light blue and dark blue default colors of the Commodore 64. (This option is not included in the other versions.) In **lines 50-74** you choose whether to work with receiver sensitivity and transmitter power in microvolts and watts or in dBm (decibels relative to 1 milliwatt). This is for your convenience only; the program will convert either type input to the other and display both as an output. In lines **100-108** you select either VHF or UHF as your operating band, and the appropriate data is read (VHF data lines **2000-2610** or UHF data lines **3000-3610**).

Actual program inputs for range calculations begin at **line 200** where you are sequentially sent to subroutines for selecting frequency, transmitter and receiver parameters, antenna gains, and antenna heights.

The subroutine in **lines 600-604** asks for a specific operating frequency within the band you selected. Actual VHF data is for 125 MHz, and UHF data is for 300 MHz, but the program scales to your actual operating frequency by a 20 × Log (F/F ref) factor in **line 350** to show propagation variations within a given band.

The subroutine in **lines 700-740** asks for receiver sensitivity and transmitter power output. The units are either microvolts and watts or dBm, depending on which you selected in **line 74**. Your input is converted to both units which will be displayed later.

The subroutine **in lines 900-906** asks for antenna gains in dBi. This is gain in dB relative to an isotropic antenna. If your antenna gain is known relative to a dipole, add 2.15 dB. For example, an antenna with 7.5 dB gain referenced to a dipole (dBd) has a gain of 9.65 dBi. You are next asked to enter any system losses, such as coaxial line losses.

Actual antenna heights are selected in the subroutine from **lines 800-870**. The data tables from lines 2000 to the end of the program contain propagation information for specific antenna heights, so you must choose a discrete value closest to your actual antenna height. For example, if your antenna height is **40** or **60** feet, use menu item 2, which is 50 feet. If you select antenna heights of 50 feet and 100 feet (menu items 2 and **3**), the program then selects H$ (2,3) data for these heights from **line 2085** or **3085** depending on whether you are operating at VHF or UHF. The string manipulation in **lines 850-858** will be described later.

The actual program output to the screen is performed **in lines 300-426**. A sample output was shown in **fig. 1A**. The operating frequency is printed, followed by XMTR output in dBm and watts and RCVR sensitivity in dBm and μv. The antenna gains and heights selected are then printed followed by the losses selected. The next item printed is the system path margin in dB, followed by the free space path loss in miles. (This is the distance over which you could communicate if it were not for the earth's curvature, useful in determining whether or not you can reach your favorite satellite.) Finally, the range over real earth is printed in both statute and nautical miles. The range given is the expected range for normal conditions; actual range will be affected by atmospheric conditions, terrain, and other factors.

One of the features of this program is that you can experiment with changes in a single parameter without having to re-enter all previous inputs. Note the menu at the bottom of the screen (see **fig. 1A**). Would you like to see how much further you could communicate if you raised your antenna from 25 to 50 feet. Just press "H" on the keyboard and you will be asked for new antenna height (subroutine at **line 800** from **line 420**), and the new range for that height will be instantly displayed. With a few keystrokes you can easily compare expected improvements in range from changes in antenna height, antenna gain, power output, and other factors. If you have a printer, just press "P" and **lines 430-436** will give you a screen dump to it.

data format

The data in the lines following **line 2000** are taken from the ESSA Technical Report mentioned in reference 2; there are about 100 pages of curves with up to 17 curves per page in that document. The programs store selected data points from various curves, and they construct "piece-wise linear" equations fitting the original curve as nearly as possible. *For the* curious, the following is a detailed explanation of the data manipulations. Assume you have selected VHF and antenna heights of 50 and 100 feet. Logical breakpoints in the ESSA curves are 10 miles for a 120 dB path, 32 miles for a 150 dB path, 50 miles for a 165 dB path, 105 miles for a 175 dB path, and 310 miles for a 210 dB path. Now look at line 2085 and note that

H$(2,3) = "120010150032165050175105210310".

The first three digits (120) store the first path point; the next three digits (010) store the first distance point; the third three digits (150) store the second path point; the fourth three digits (032) store the second distance point, and so on. The last three digits (310) represent the last mileage point. After antenna heights are chosen (**lines 800-832**), the appropriate data line is divided up into five path points and five distance points by string manipulation in **lines 840-858**.

After the program calculates your system path margin (**lines 320** and **350**) from the various inputs, the program path is compared to the data points described

above. If it is less than the smallest or greater than the largest, a "Range Not In Program" message is printed (**lines 352-358**). If the path margin falls between the data endpoints, the program calculates expected range by assuming a straight line between the nearest two points stored (**lines 360-366**), and the expected range is printed to the screen.

entering the program

Enter the program as listed, taking the normal precautions to SAVE it before you RUN it so that if you make a typing error that could cause a lock-up, you'll be able to go back to the saved version without having to retype the entire program.

As listed, the program contains data statements for altitudes to 100,000 feet (includes air-to-ground and air-to-air data for you aeronautical mobile enthusiasts.) If you plan to use the program only for ground-to-ground communications, you can omit any data statements with H$(x,y) array subscripts greater than 4. For example, for VHF data including antenna heights of 25, 50, 100, and 500 feet, you need to type only **lines 2000-2020, 2080-2090, 2150-2155, 2215,** and **2610**. Just don't try to select a height (**lines 800-832**) for which you didn't enter any data.

If you don't want to keystroke the Commodore-64 program yourself, send a blank tape or formatted disk with a stamped, self-addressed tape or disk mailer, and a check or money order for $5.00 to Lynn A. Gerig, R. R. #1, Monroeville, Indiana 46773, and two verified copies will be made for you. A similar program — ground-to-ground only, with some other features missing — that has been "crunched" to fit within the 3.5K memory of an unexpanded VIC-20 is available under the same terms. (A printed listing of any programs described is available for $1 and a self-addressed envelope with two first-class stamps attached.)

TI-99/4A program

Except for **lines 22-40**, which poke screen and letter colors to appropriate memory locations, the Commodore 64 program described above has nothing particularly unique to that machine, and can easily be converted to run on other computers using BASIC. Subtle differences between machines, however, make certain conversions necessary. For example, although the Commodore clears the screen with a "PRINT CHR$(147)", some other brands use "CALL CLEAR" or some other command. In addition, some computers will not permit the use of multiple statements on a single BASIC line number.

A program for the TI-99/4A is listed in **fig. 2.** Because this computer has only 16K of memory, and the complete program requires about 20K of RAM, separate VHF and UHF programs are required, and not

fig. 2. Communications range calculation program for the TI-99/4A.

```
10 GOSUB 222
12 CALL CLEAR
14 PRINT TAB(3);"VHF PROPAGATION PROGRAM"
14 PRINT
18 PRINT TAB(7);"FOR THE TI-99/4A"
20 PRINT
22 PRINT TAB(14);"BY"
24 PRINT
26 PRINT TAB(9);"J.R. HENNEL"
28 FOR DELAY=1 TO 1000
30 NEXT DELAY
32 GOTO 34
34 CALL CLEAR
36 PRINT "THIS PROGRAM WILL CALCULATE"
38 PRINT "EXPECTED RANGES FOR VHF(100"
40 PRINT "-175 MHZ) FREQUENCIES.  AP-"
42 PRINT "PROXIMATE DYNAMIC RANGE IS"
44 PRINT "FOR PATH LOSSES OF 125 TO"
46 PRINT "200 DB, COVERING MOST AP-"
48 PRINT "PLICATIONS FOR RCVR SENSI-"
50 PRINT "TIVITIES FROM .5 TO 10 MIC-"
52 PRINT "ROVOLT AND XMTR PWR OF 1 TO"
54 PRINT "1000 WATTS.  THE PROGRAM "
56 PRINT "COVERS ANTENNA HEIGHTS FROM"
58 PRINT "25 FT TO 40,000 FT. PROGRAM"
60 PRINT "DEFAULTS TO RCVR SENS AND"
62 PRINT "XMTR PWR IN DBM. WOULD YOU"
64 PRINT "RATHER WORK WITH MICROVOLTS"
66 PRINT "AND WATTS? (Y=YES)"
68 INPUT D$
70 GOSUB 2005
72 GOSUB 250
74 GOTO 76
76 PRINT
78 GOSUB 266
80 PRINT
82 GOSUB 400
84 PRINT
86 GOSUB 328
88 PRINT
90 CALL CLEAR
92 PRINT " VHF PROPAGATION:"
94 PRINT TAB(3);" FREQUENCY   ";F;TAB(24);"MHZ"
96 PRINT " TRANSMITTER POWER OUT:"
98 PRINT TAB(3);"=";PD;TAB(10);"DBM =";PW;TAB(24);"WATTS"
100 PRINT " RECEIVER SENSITIVITY:"
102 PRINT TAB(3);"=";RD;TAB(10);"DBM =";RM;TAB(24);"UV"
104 PRINT " LOWER ANTENNA:"
106 PRINT TAB(3);"=";GL;TAB(10);"DBI @ ";H1;TAB(24);"FT"
108 PRINT " UPPER ANTENNA:"
110 PRINT TAB(3);"=";GU;TAB(10);"DBI @ ";H2;TAB(24);"FT"
112 PRINT " COAXIAL LINE LOSSES:"
114 PRINT TAB(3);"=";LL;TAB(10);"DB"
116 PL=PD-RD+GU+GL-LL
118 DF=PL-37-20*LOG(F)/LOG(10)
120 DF=10^(DF/20)
122 DF=INT(DF+.5)
124 PRINT TAB(3);PL;"DB";TAB(14);"PATH"
126 PRINT " FREE SPACE PATH=";DF;TAB(24);"MILES"
128 PL=PL-20*LOG(F/F1)/LOG(10)
130 PRINT
132 IF PL<P1 THEN 134 ELSE 136
134 PRINT "RANGE NOT IN PROGRAM; <";D1;"MILES"
136 IF PL<P1 THEN 138 ELSE 142
138 PRINT TAB(23);"<";D1;"(NAUT MI)"
140 GOTO 176
142 IF PL>P5 THEN 144 ELSE 146
144 PRINT "RANGE NOT IN PROGRAM; >";D5;"MILES"
146 IF PL>P5 THEN 148 ELSE 152
148 PRINT TAB(23);">";D5;"(NAUT MI)"
150 GOTO 176
152 IF (PL>=P1)*(PL<P2)THEN 154 ELSE 156
154 DN=D1+(D2-D1)*(PL-P1)/(P2-P1)
156 IF (PL>=P2)*(PL<P3)THEN 158 ELSE 160
158 DN=D2+(D3-D2)*(PL-P2)/(P3-P2)
160 IF (PL>=P3)*(PL<P4)THEN 162 ELSE 164
162 DN=D3+(D4-D3)*(PL-P3)/(P4-P3)
164 IF (PL>=P4)*(PL<P5)THEN 166 ELSE 168
166 DN=D4+(D5-D4)*(PL-P4)/(P5-R4)
168 DS=INT(DN*1.151+.5)
170 DN=INT(DN+.5)
172 PRINT " MAXIMUM EXPECTED RANGE:"
174 PRINT TAB(3);"=";DS;TAB(12);"MILES =";DN;TAB(24);"N MI"
176 PRINT LL$
178 PRINT " G=MOD ANT GAIN  R=RUN AGAIN"
180 PRINT " H=MOD ANT HT     F=NEW FREQ"
182 PRINT " X=MOD R/T SENS/PWR  Q=QUIT"
184 INPUT A$
186 IF A$=" " THEN 184
188 PRINT
190 PRINT
192 IF A$="R" THEN 72 ELSE 194
194 IF A$="Q" THEN 196 ELSE 198
196 END
198 IF A$="G" THEN 200 ELSE 204
200 GOSUB 400
202 GOTO 90
204 IF A$="H" THEN 206 ELSE 210
206 GOSUB 338
208 GOTO 90
210 IF A$="X" THEN 212 ELSE 216
212 GOSUB 266
214 GOTO 90
216 IF A$="F" THEN 218
218 GOSUB 250
220 GOTO 90
222 CALL CLEAR
224 D$="N"
226 DIM H$(12,12)
228 DIM H(12)
230 H(1)=25
232 H(2)=50
234 H(3)=100
236 H(4)=500
238 H(5)=1000
240 H(6)=2000
```

```
242 H(7)=5000
244 H(8)=10000
246 H(9)=15000
248 H(10)=20000
250 H(11)=30000
252 H(12)=40000
254 LL$="----------------------------"
256 RETURN
258 PRINT " FREQUENCY IN MHZ (100-175)"
260 INPUT "       ";F
262 IF (F<FL)+(F>FU)THEN 258
264 RETURN
266 IF D$="Y" THEN 268 ELSE 272
268 INPUT "INPUT XMTR PWR (IN WATTS)          ";PW
270 GOTO 292
272 INPUT "INPUT XMTR POWER (IN DBM):         ";PD
274 PRINT
276 PW=(PD-30)/10
278 PW=10^PW
280 IF PW>=1 THEN 282 ELSE 286
282 PW=INT(PW*10+.5)/10
284 GOTO 298
286 IF PW<1 THEN 288
288 PW=INT(PW*100+.5)/100
290 GOTO 298
292 PD=10*LOG(PW)/LOG(10)+30
294 PD=INT(PD*10+.5)/10
296 PRINT
298 IF D$="Y" THEN 300 ELSE 304
300 INPUT " RCVR SENS (IN MICROVOLTS)         ";RM
302 GOTO 322
304 INPUT "RCVR SENSITIVITY (IN DBM)         ";RD
306 RM=(RD+107)/20
308 RM=10^RM
310 IF RM>=1 THEN 312 ELSE 316
312 RM=INT(RM*10+.5)/10
314 GOTO 326
316 IF RM<1 GOTO 318
318 RM=INT(RM*100+.5)/100
320 GOTO 326
322 RD=20*LOG(RM)/LOG(10)-107
324 RD=INT(RD*10+.5)/10
326 RETURN
328 REM  ANTENNA HEIGHTS
330 CALL CLEAR
332 PRINT "CHOOSE ANTENN HEIGHTS BY"
334 PRINT "SELECTING NUMBER FROM THE"
336 PRINT "FOLLOWING MENU"
338 PRINT
340 PRINT
342 PRINT "1=  25' 5=  1000'  9= 15000'"
344 PRINT "2=  50' 6=  2000 '10= 20000'"
346 PRINT "3= 100' 7=  5000' 11= 30000'"
348 PRINT "4= 300' 8= 10000' 12= 40000'"
350 PRINT
352 PRINT
354 INPUT "SELECT HEIGHT OF LWR ANTENNA:   ";H1
356 PRINT
358 INPUT "SELECT HEIGHT OF UPR ANTENNA    ";H2
360 PRINT
362 IF (H1<1)+(H2<1)+(H1>12)+(H2>12)THEN 364 ELSE 368
364 PRINT "NOT IN MENU"
366 GOTO 354
368 IF H1>H2 THEN 370 ELSE 374
370 PRINT " LOWER AND UPPPER REVERSED"
372 GOTO 354
374 HC$=H$(H1,H2)
376 H1=H(H1)
378 H2=H(H2)
380 P1=VAL(SEG$(HC$,1,3))
382 D1=VAL(SEG$(HC$,4,3))
384 P2=VAL(SEG$(HC$,7,3))
386 D2=VAL(SEG$(HC$,10,3))
388 P3=VAL(SEG$(HC$,13,3))
390 D3=VAL(SEG$(HC$,16,3))
392 P4=VAL(SEG$(HC$,19,3))
394 D4=VAL(SEG$(HC$,22,3))
396 P5=VAL(SEG$(HC$,25,3))
398 D5=VAL(SEG$(HC$,28,3))
400 S1=INT(D1*1.151+.5)
402 S5=INT(D5*1.151+.5)
404 CALL CLEAR
406 RETURN
408 PRINT " GAIN OF LWR ANT (IN DBI):"
410 INPUT "       ";GL
412 PRINT
414 PRINT " GAIN OF UPR ANT (IN DBI):"
416 INPUT "       ";GU
418 PRINT
420 INPUT "COAXIAL LINE LOSSES (IN DB)        ";LL
422 RETURN
424 F1=125
426 FL=100
428 FU=175
430 REM  VHF DATA
```

all the altitudes used in the Commodore program will fit.

To devise a VHF program for your TI-99/4A, enter the program listed in **fig. 2**. At the end of program, type the VHF data statements (**lines 2005-2610**) listed in the Commodore version (**fig. 1**). Since the data statements are the same for all versions, they are not listed again in **fig. 2**. However, because the TI memory is too small for the complete program, only altitudes to 40,000 feet should be included. *Do not enter data lines with H$(x,y) subscripts of 13, 14, or 15. For example, do not enter* lines **2065-2075**, **2135-2145**, etc. Refer to the previous section if you want only a short ground-to-ground version.

For a UHF program for your TI, use the program listed in **fig. 2**, making the following changes:

- In lines **14**, **34**, and **430**, change "VHF" to "UHF."
- In lines **38** and **426**, change "100" to "200."
- In lines **40** and **428**, change "175" to "500."
- In line **70**, change "GOSUB 2005" to "GOSUB 3005."
- In line **258**, change "100-175" to "200-500."
- In line **424**, change "125" to "300."

Now add the UHF data statements (**lines 3005-3610**) listed in the Commodore version (**fig. 1**), again deleting the data above 40,000 feet. Again for this program, only those data statements with H$(X,Y) subscripts of 1, 2, 3, and 4 need to be entered if you are interested in ground-to-ground calculations only.

If you don't want to type the TI programs yourself, send a blank tape — no disks — with stamped, self-addressed tape mailer, and a check or money order for $5.00 to Joe Hennel, 4316 Winston Drive, Ft. Wayne, Indiana 46806.

using the programs

The program must be used with caution. Each provides ranges over average terrain for which communications can be expected 50 percent of the time. Obviously you won't be able to communicate as far through dense jungle or through mountainous areas, so some common sense must be used. However, the programs are very useful for determining relative changes in anticipated range due to modifications to receivers, transmitters, and antennas.

When entering receiver and transmitter parameters, use power output (not input) at the transmitting end, and sensitivity at the receiving end. For example, if you are running a full kW, and your friend is running only 10 watts, he may be able to hear you without your being able to hear him.

Feel free to contact either of the authors — at their addresses given at the opening of this article — with questions or comments about the programs; only letters with an SASE enclosed will be answered.

references

1. *Propagation Curves, Report 966-6C,* National Defense Research Committee 15, Bell Telephone Laboratories, Inc., Issue 3, October, 1944 (declassified to OPEN status March 8, 1946).

2. *ESSA Technical Report ERL 111-ITS 79, Transmission Loss Atlas for Select Service Bands from 0.125 to 15.5 GHz,* Institute for Telecommunication Sciences, Boulder Colorado, May, 1969. (Available for $1.25 from Superintendent of Documents, United States Government Printing Office, Washington, D. C. 20402.)

ham radio

a 432-MHz, 1500-watt amplifier

An 8938 triode in a coaxial cavity provides maximum output power

This article describes an amplifier that uses an 8938 coaxial triode in a commercially available cavity assembly, and is conservatively rated for CW and SSB operation at 1500 watts output on 432 MHz. A driver capable of 100 watts output is required. With proper cavity adjustment, efficiency of over 50 percent and power gain of 12 dB are readily obtainable at high output levels. Construction is straight-forward.

The complete amplifier assembly (**fig. 1**) consists of three units: the RF section, the metering and control unit and the power supply. These units are interconnected by cables, using MHV type connectors and RG-59 cable for the high voltage and C-J (Cinch Jones) connectors with appropriate low voltage cables. The heater wires in the low voltage cables must use a conductor large enough to provide at least 4.55 volts at the cavity terminals.

This type of construction provides maximum flexibility. Each unit can be located in the most favorable position for its particular function, thereby simplifying maintenance.

RF section

The cavity assembly (**fig. 2**) is mounted on a chassis measuring 10 inches (25.4 cm) wide, 17 inches (43 cm) long and 5 inches (12.7 cm) high, supported by metal stand-offs at the four corners of the upper plate of the cavity. A square opening to match the size of the EMI filter mounted on the bottom of the cathode cavity is cut into the chassis. Holes to match the four mounting holes for the EMI filter are drilled through the

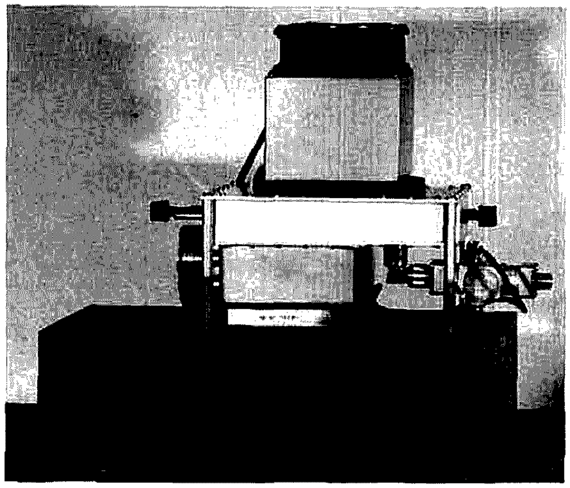


fig. 1. Front view of amplifier assembly.

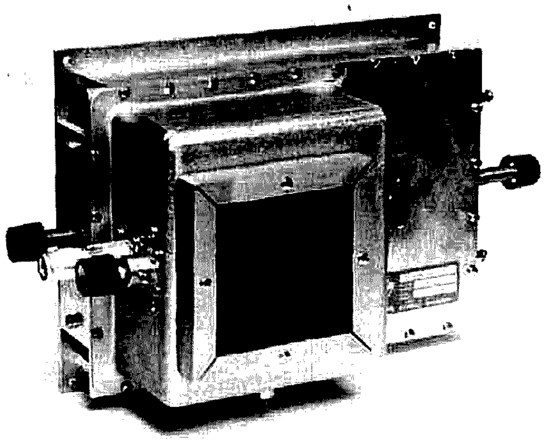


fig. 2. Cavity assembly (*photo courtesy of EIMAC Varian*).

By **F. J. Merry, W2GN**, P. O. Box 546, 35 Highland Drive, East Greenbush, New York 12061

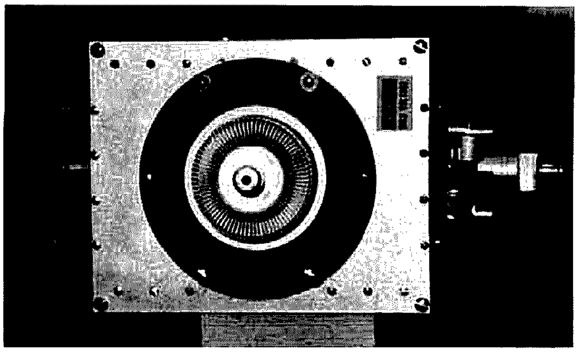


fig. 3. Top view of cavity with tube installed. The screws with washers are the support screws for the top cabinet. The double line section for measuring forward and reverse RF power output is visible on the right.

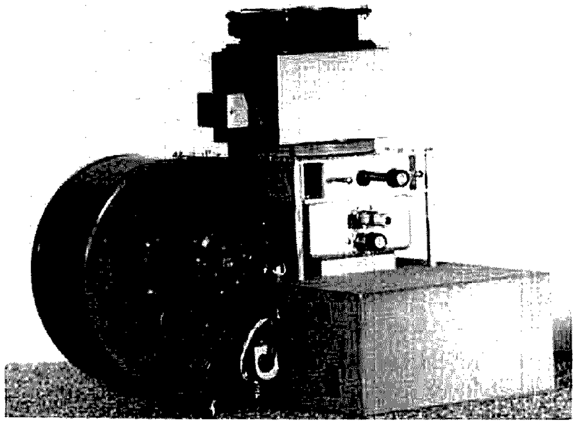


fig. 4. Rear and left side view. Air switch mounted on blower is visible. Also blower connections. The upper control on the cavity is the tuning control. Below it is the RF input and load control. Next below is the cathode tuning control.

chassis (**fig. 3**). The cavity is thus mounted to the chassis by four screws into the standoffs at the four corners of the cavity top plate and four screws coming up through the bottom of the chassis and through the EMI filter.

The blower (265 CFM) is attached to a mounting plate (**figs. 4, 5**) fastened to the rear of the chassis. A hole to match the blower output size is cut into the mounting plate and the chassis. No screen is required over the chassis air input since the EMI filter performs this function. An air switch is mounted in the output air stream of the blower.

An auxiliary blower (55 CFM), shown in **fig. 6**, is mounted on top of a small cabinet measuring 4 × 5 × 6 inches (10.2 × 12.7 × 15.2 cm). The bottom plate of this cabinet is not used. The cabinet is mounted on top of the insulating ring of the cavity using four of the six insulating ring mounting screws. These four mounting screws are replaced with slightly longer screws to permit a secure mounting while avoiding any protrusion into the cavity. A hole is punched in the top cover of the cabinet to match the fan blade diameter. The blower is mounted to the top cover by securing it with adhesive to felt strips around the periphery of the blower. These felt strips have an adhesive backing that provides secure fastening to the cabinet top. (Care must be exercised to trim the strips so that the fan blade will not catch on the felt.) The blower is mounted, of course, so that the air stream flows upward.

The cabinet discussed protects users from the high voltage to the anode of the 8938. The MHV type connector is mounted on a small box on the rear of the cabinet. The 0.001/4 kV feedthrough capacitor and a small screw fasten the small box to the cabinet. The feedthrough capacitor is positioned slightly above the level of the tube anode. A five-turn, 1/4-inch diameter (1-inch long) RFC is connected between the capacitor and the anode connecting clip.

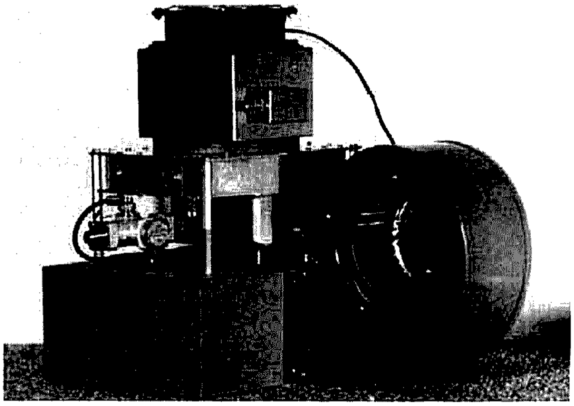


fig. 5. Rear and right side view. The double line section is visible, attached to the cavity RF output connector. Above it is the cavity load control. Just in front of the large blower are the heater cathode terminals on feedthrough capacitors. The high voltage MHV connector mounted on the small box on the upper cabinet is also in view.

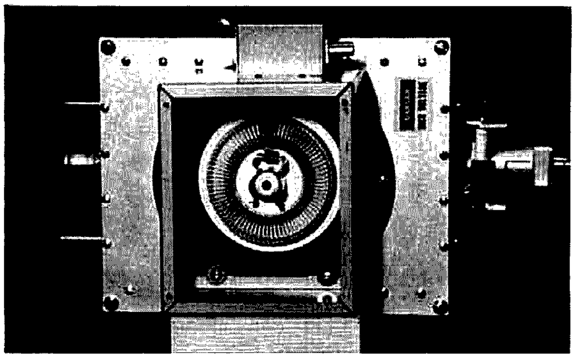


fig. 6. The top view with the auxiliary blower removed shows the HV connection to the tube through an RF choke. Also shown is the method of mounting the small upper chassis box to the cavity. (Don't be confused by the wood block used to support the cavity for the picture.)

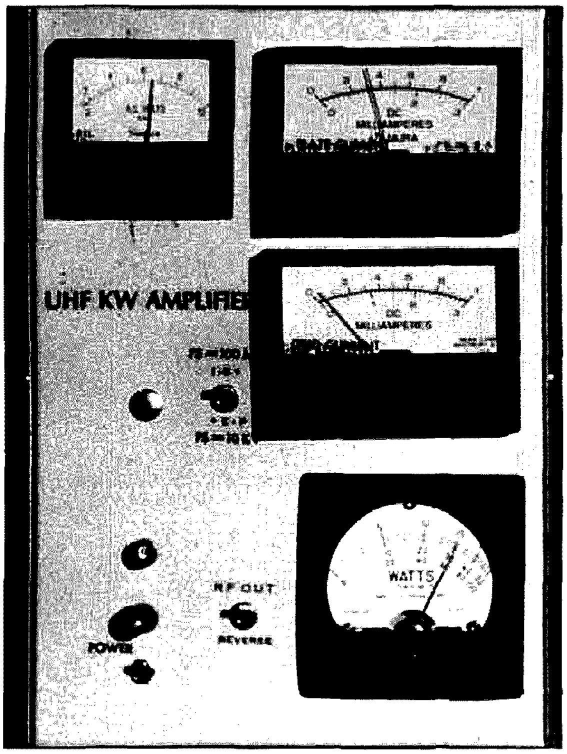


fig. 7. Front view of control unit. Note that the amplifier is in operation with 1500 watts output at 1-ampere plate current and 6-mA grid current.

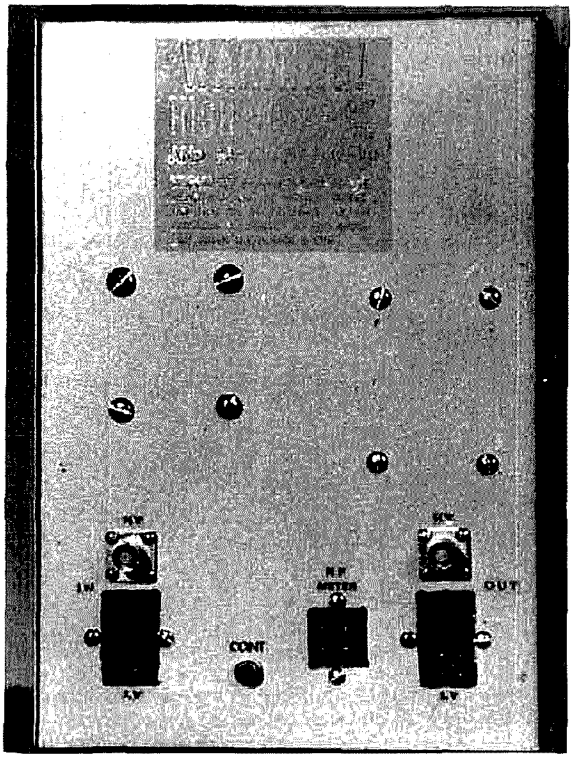


fig. 8. Rear view of control unit.

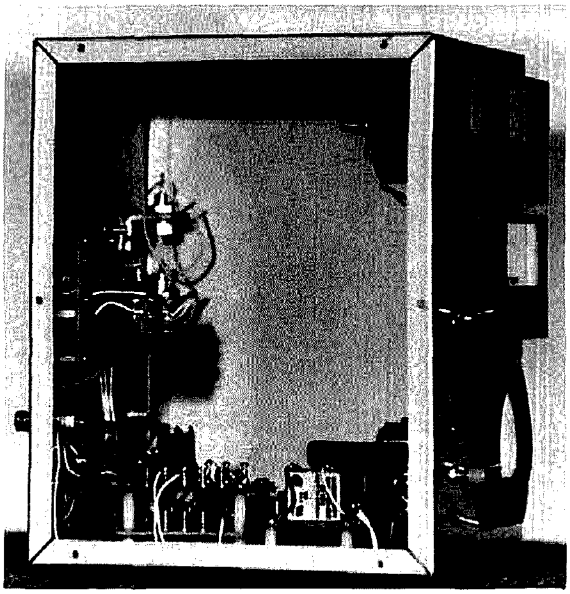


fig. 9. Left side of control unit with cover off.

Three C-J connectors are mounted on the rear of the chassis (**fig. 4**). The cable from the blower is shown plugged into a four-conductor C-J connector. This cable furnishes 120 VAC to the blower motor and provides connection to the air switch. The 120 VAC is bridged to a two-conductor C-J connector to furnish power to the auxiliary blower. The other four-conductor C-J connector, to the right of the blower connector, connects to the RF power output forward and reverse RF meter elements of the double line section, which is connected to the RF output connector on the cavity. The output connector on the cavity is of the HN type. A right-angle HN adapter connects and supports the double line section, which is equipped with an HN type QC connector. The RF output connector on the line section may be either HN or N, with HN recommended. The eight-conductor low voltage connector is also visible on the rear of the chassis. The interior of the chassis is vacant except for the wiring to the C-J connectors. The small openings in the chassis at the corners are sealed with plastic tape to prevent air leakage.

control unit

The control unit (**figs. 7, 8**) is assembled in a cabinet measuring 7 × 8 × 10 inches (17.8 × 20.3 × 25.4 cm). The circuitry is divided into sections mounted on individual terminal boards as follows:

- metering and stand-by bias switching control on a two-section board;
- high voltage failure protection on two single-section boards, one mounted over the other;

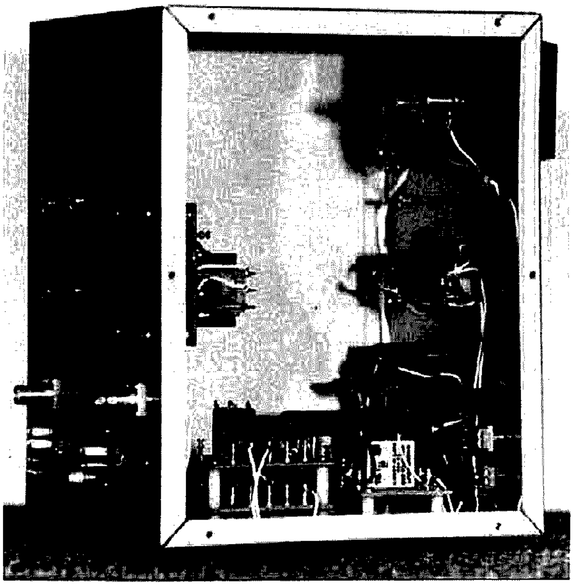


fig. 10. Left side of control unit view from rear.

• 12 VDC supply on a single board.

The two 50-watt zeners are mounted on brackets secured to the rear of the cabinet. One of the zeners is visible in **fig. 9** and **10**. The other is obscured behind

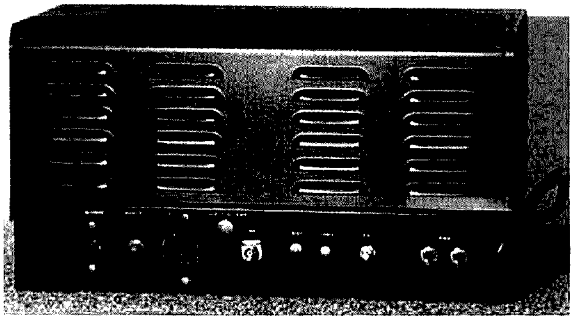


fig. 11. Power supply. The control locations marked BIAS, CONT and SCR are not used in this application. The BLOWER outlet is also not used since the two blowers on the amplifier chassis are powered over the cable connector marked LV.

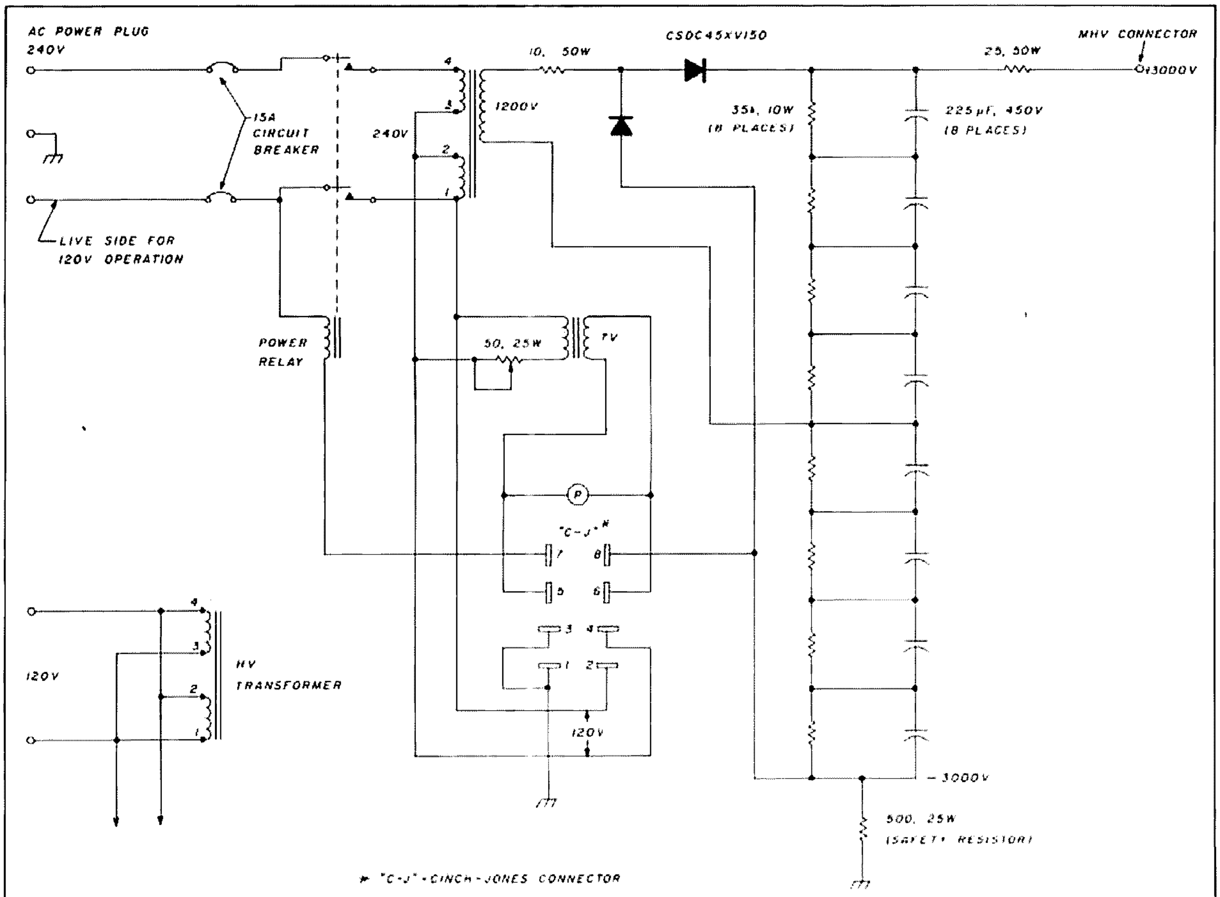


fig. 12A. Power supply schematic.

the 12 VDC board. Two 13-volt zeners in series provide the 26-volt cathode bias.

The mounting of the meters and switches on the front panel and the connectors on the back may be observed from the photos. Meter scales are as follows:

- plate current 0-3 amperes
- grid current 100 milliamperes
- plate voltage 10 kilovolts (read plate voltage by depressing non-locking switch next to the meter)
- heater voltage 0-10 volts AC
- RF output 2500 watts forward, 250 watts reverse (read reverse by depressing non-locking switch next to the meter)

The small toggle switch on the front panel at the lower right (**fig. 11**) prepares the circuitry for operation of the power relay in the power supply. When depressed, the non-locking push-button switch above the power switch causes the power relay to energize the power supply. As the blower comes up to speed, the air switch engages and locks the power relay up after the release of the push-button switch. The amplifier is thus protected from air supply failure.

Note: the schematic (**fig. 12B**) shows the transistor connected through a 15-megohm resistor to the high voltage line. When high voltage is present, this transistor conducts, activating its associated relay. When this relay is energized, operating bias is applied to the amplifier through another relay which is powered on by a closure to ground of the control jack during transmit. The high voltage portion of this circuitry protects the 8938 tube in case excitation is applied and high voltage is not on.

The metering and other protective features are standard for a grounded grid triode amplifier and may be observed by analyzing the schematic.

power supply

The power supply (**figs. 11** and **12A**) is capable of providing 1 ampere at about 3000 volts CCS (continuous commercial service). The voltage doubling circuit, with a net of about 25 μF of filter capacity, has three protective resistors: 10-ohm 50-watt diode protection,

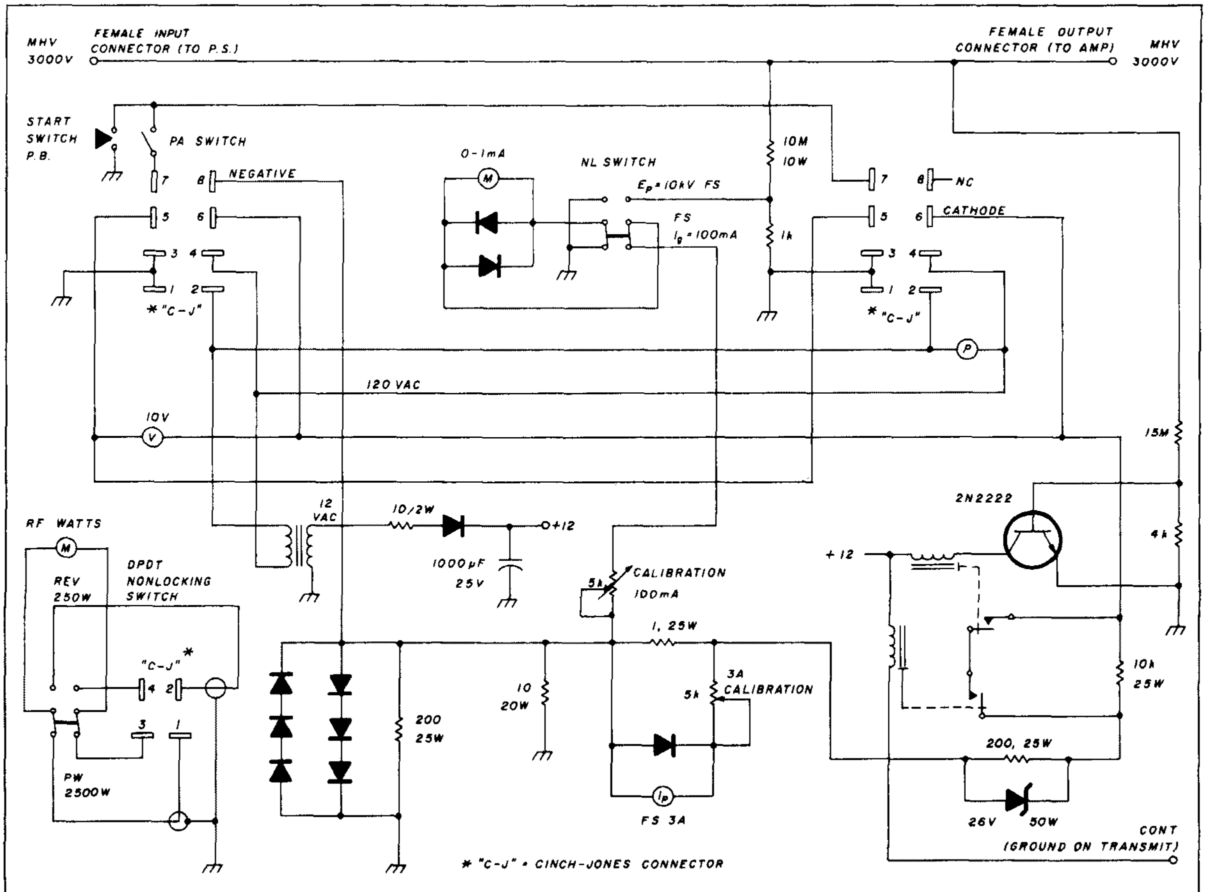


fig. 12B. Control and metering schematic.

25-ohm 50-watt output short circuit protection, and a 500-ohm 25-watt from negative to chassis for safety. The rest of the circuitry can be easily determined from a review of the schematic.

Power supply parts are mounted in a 17 × 10 × 3 inch (approximately 43 × 25 × 7 cm) steel chassis with a ventilated cover. The transformer should have a 3 kVA rating ICAS (intermittent continuous Amateur service). The unit shown in **fig. 11** weighs a bit over 50 pounds, or 22.68 kilograms.

operation

To power up the amplifier, place the two power supply circuit breakers in the on position. Place the power switch on the control unit in the on position. Push the non-locking push-button switch on the control unit in and hold until the blower comes up to speed. Release the push-button switch and allow a minimum of three minutes warm-up time. Check plate voltage by placing the switch near the grid current meter in its non-locking position. Observe the heater voltage. Because of voltage drop between the control unit and the cavity, it will read (typically), 6.5 VAC when the voltage at the cavity terminals is 4.6 The reading of the heater voltage at the control unit will, obviously, vary with the length of cable between the two units. This irregularity can be avoided by providing a pair of wires in the cable to connect the cavity terminals directly to the voltmeter (see **fig. 12C**).

The amplifier requires a driver capable of providing 100 watts, a control cable, and a dummy load that can handle 1500 watts at this frequency. The amplifier is now ready to be tested.

After warm-up, apply the control signal (ground) but do not apply drive to the amplifier. Observe an idling current of about 0.125 ampere. With a drive power of 10 to 15 watts, adjust the cathode tune and load controls for a rise in plate current. Do not exceed 0.5 to 0.75 ampere of plate current during preliminary tests. Next, adjust the plate cavity tune and load con-

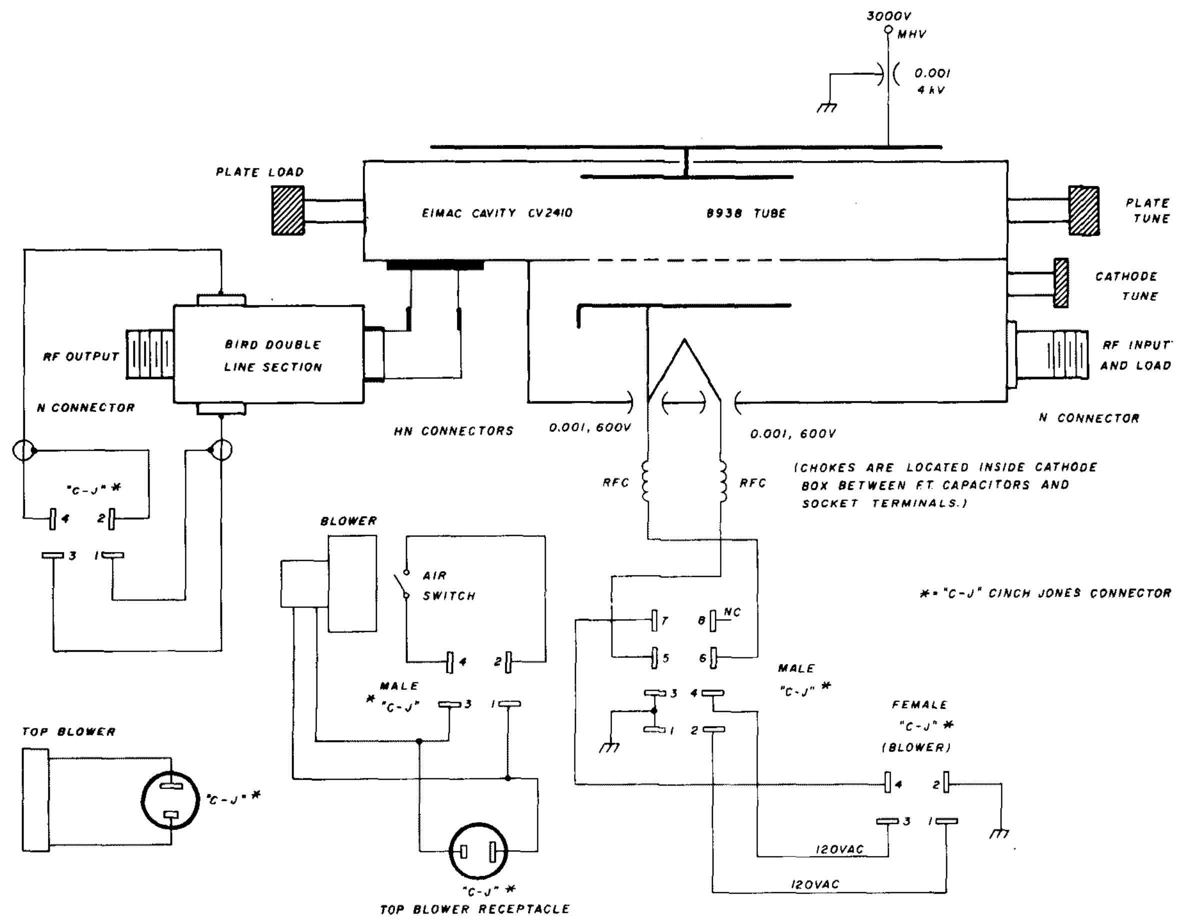


fig. 12C. Cavity and blower schematic.

trols for an indication of output power. From this point on, the usual format of amplifier adjustment is followed. Optimize the load and tune controls as the drive power is increased until 1500 watts of output is obtained at nominal plate current. The reverse RF power on the drive side of the amplifier will be close to zero. Typical readings at 1500 watts output are as follows:

- drive power, 100 watts
- plate voltage, 2800
- plate current, 1.00 ampere
- grid current, 6 milliamperes

With these readings, the efficiency is 54 percent and the power gain is about 12 dB.

It is suggested that adjustments to the plate cavity be made in steps with the drive power *off*. The mounting of the RF unit should be done in such a manner that the plate cavity cannot be inadvertently contacted by the body during operation.

The tests on this amplifier had to be limited to a maximum of 2 minutes (key down) because my dummy load was rated at one kW dissipation. However, a few very short "shots" of higher drive power showed outputs of up to 2500 watts. Dick Frey, W2SZ/1, super-VHF/UHF contester of Mount Greylock, Massachusetts, said, during one of the tests, "Man, there's an amplifier that sez *'gimme more!'*"

ham radio

micros and VHF beacons transmit messages automatically

Tied to beacons, micros find best times to dump data and predict best times for data transmission

Like many other Amateurs, I purchased a home computer several years ago. I started with a VIC-20 and now have several VIC-20 and Commodore 64 computers. My machines have been used for many applications, including several in the hamshack.

For high-frequency work, the availability of Bob Rose and Associates' "MINIMUF" program has been a blessing.[1] Many others have written improvements to the basic algorithm that make it even more versatile. But except for some simple programs for calculating repeater coverage and such, the use of computers in VHF propagation studies has been largely ignored.

radio link for program exchange

My brother Jon, WB9YJC, an electrical engineering student, has been bitten by the computer bug too, and uses both a VIC-20 and Atari 400 in his Amateur activities. Because we often work on developing programs together, but live 70 miles (112 km) apart, we found program exchange by mail much too slow, and telephone communication too expensive. A 2-meter data link seemed a likely way to solve the problem of keeping in touch.

Here in the Midwest, this would seem to be a reasonable solution. But other considerations — particularly antenna height limitations — can make the 2-meter data link more difficult than it might initially appear to be.

```
100 OPEN 2,2,3,CHR$(128+35):T7=37136:T9=37138:PRINT"J"
110 PRINT"ENTER TIME":INPUT TI$:PRINT "J"
120 A$="":GOTO 200
130 POKE T9,PEEK(T9)OR32:POKE T7,PEEK(T7)AND32
140 FOR X=1TO1000:NEXT
150 PRINT#2," DE KZ9XYZ BEACON SPRINGFIELD,";
155 PRINT#2," ILL TRANSMITTING FOR PROPAGATION STUDY";
160 PRINT#2," U*U*U*U*U*U*U*U*U*U*U*U*U*U*";
165 PRINT#2," U*U*U*U*U*U*U*U*U*U*U*U*U*U*";
170 PRINT#2," DE KZ9XYZ BEACON SPRINGFIELD ILL AR"
180 FOR Z=1TO12000:NEXT Z
190 POKE T9,6:GOTO120
200 GET#2,A$:IF A$=""THEN 200
210 PRINT TI$,
220 GET#2,A$:IF A$<>""THEN 220
230 GOTO 130
240 REM ***************
250 REM *             *
260 REM * "KERCHUNK"  *
270 REM * (C) 1985    *
280 REM * JIM GRUBBS  *
290 REM * PO BOX 3042 *
300 REM * SPRINGFIELD *
310 REM * IL 62708    *
320 REM *             *
330 REM ***************

READY.
```

fig. 1. This program listing for either the VIC-20 or the C-64 implements the Kerchunk program explained in the text. It is designed to be used with a Kantronics interface. You can substitute your own message in lines 150, 160, and 170.

Both of us have busy schedules, and our free times seldom coincide. Before we could use a radio link for program exchange, we would first need to establish that a predictable, everyday path would be possible, and then determine the best times and conditions for transmission.

The first breakthrough came when Jon suggested installing a computer-controlled beacon at his QTH. A simple CW program with some timing subroutines for his VIC-20 soon placed WB9YJC/Beacon on the air. A modification of one of my own programs soon put my signal on the air as well, and although the preliminary results were not very good, we at least had a "semi-reliable" signal for which we could listen.

I say "semi-reliable" because with Jon's VIC-20 tied up as a beacon controller, he was unable to use it for any other purpose. His Atari 400, on the other hand, was seldom used. To take advantage of its availability,

By Jim Grubbs, K9EI, P.O. Box 3042, Springfield, Illinois 62708

an Atari version of the Morse-code beacon was developed and the VIC-20 was freed for other uses.

As time permitted, we were then able to observe the path between us anytime, day or night. With only 10 watts on each end and a directional antenna on only one end of the path (his end), we were not very successful. It was obvious that improvements would be necessary.

While I contemplated additional power and an improved antenna system, Jon raised his antenna by 10 feet (3.05 meters). His signal was now up noticeably.

automatic signal logging

In the absence of an operator on either end to confirm reception, the effectiveness of our beacons was still limited. Some sort of logging system needed to be devised, at least on one end, to record times of reception when on operator was present.

My station consists of a VIC-20, a Kantronics interface, and a 10-watt transceiver. By having the VIC-20 constantly searching out input from the Kantronics interface — i.e., an incoming signal — a crude form of carrier operated (actually a voice or noise operated) relay could be built, using software.

When the squelch opened on an incoming signal, audio was fed to the interface. White noise coming through a terminal unit fortunately appears to the computer as random ASCII characters. By using a simple GET loop from the RS-232 port, the presence of an incoming signal could be noted. Using the TI clock in the VIC-20, the time of reception could then be logged directly on the screen or sent to a printer.

With my station operating in the "kerchunk" mode, it was now possible for me to know whenever my system "heard" an incoming signal or, at least, noise. Each time *it heard an* incoming signal, it would respond with a brief ASCII message; partial automation, with some *unmanned vertification* of results, had been achieved. But just as repeater kerchunking may invite abuse, so might this type of operation. A more complete solution was obviously required.

autorespond program

Jon used his knowledge of assembly language to design an ASCII program for the Atari 400. Not having an assembler for the Atari, he assembled the *program* on the VIC-20 and then transferred it to the Atari, a neat trick possible because both machines are built around the 6502 processor. The WB9YJC/Beacon was now transmitting in full ASCII, using an AEA CP-1 interface.

Meanwhile, I was developing a positive autorespond program — in reality, just an improvement on the earlier monitoring program. Borrowing from an input routine I designed for a landline-based bulletin board program, the program now constantly listens for a specific sequence of characters; old RTTY operators will recognize it as SELCAL.

In operation, the reception of a carriage return clears the memory. The next four characters are then re-

```
100 OPEN 2,2,3,CHR$(128+35):T7=37136:T9=37138:PRINT"J"
110 PRINT"ENTER TIME":INPUT TI$:PRINT "J"
120 A$="":GOTO  200
130 POKE T9,PEEK(T9)OR32:POKE T7,PEEK(T7)AND32
140 FOR X=1TO1000:NEXT
150 PRINT#2," DE KZ9XYZ BEACON SPRINGFIELD, ILL",
155 PRINT#2," TRANSMITTING FOR PROPAGATION STUDY";
160 PRINT#2," U*U*U*U*U*U*U*U*U*U*U*U*U*";
165 PRINT#2," U*U*U*U*U*U*U*U*U*U*U*U*U*";
170 PRINT#2," DE KZ9XYZ BEACON SPRINGFIELD ILL AR"
180 FOR Z=1TO12000:NEXT Z
190 POKE T9,6:GOTO120
200 LN=4:GOSUB 230
210 IF A$="NNNN" THENPRINT TI$,:FORO=1TO3000:NEXT:GOTO 130
220 GOTO 200
230 A$="":Y$=""
240 GET#2,Y$
250 IF Y$="" THEN GOTO 240
260 IF Y$=CHR$(13) THEN RETURN
270 IF Y$<" " OR A$>"Z" THEN GOTO240
280 A$=A$+Y$:IF LEN(A$)>LN THEN RETURN
290 GOTO 240
300 REM ***************
310 REM *             *
320 REM * AUTORESPOND *
330 REM * (C) 1985    *
340 REM * JIM GRUBBS  *
350 REM * PO BOX 3042 *
360 REM * SPRINGFIELD *
370 REM * IL 62708    *
380 REM *             *
390 REM ***************

READY.
```

VIC-20 or Commodore-64 computer and Kantronics interface will respond to a preset turn-on code. The code is located in line 210 (NNNN in the example). It can be any four letter code of your choosing. It must begin and end with a carriage return when sent by the distant station. The response message in lines 150, 160, and 170 can be changed to suit your needs.

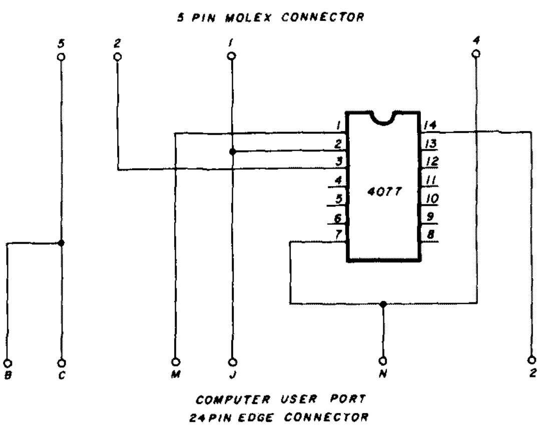


fig. 3. Interface circuit to connect your Commodore computer to the Kantronics interface. It requires a single 4077 integrated circuit and connects to the 24-pin user port located on the rear of your computer. Power for the IC is supplied from this connector also.

ceived and matched to the SELCAL being used, unless another carriage return is received. If another carriage return is received, the program simply starts over again. The sequence must be terminated by a carriage return in order for the match to take place. The addition of a simple sequence on the end of Jon's beacon message causes my station to autorespond.

Just as in "kerchunk," autorespond logs the activation time to the screen. Now, of course, there is reasonable certainty that the activation is caused by the signal at the other end of the path, rather than random noise or other signals.

It is a simple matter to include the time in the response message. The station on the other end can then check the screen or printer for the presence or absence of a response, with the time duly noted.

The sound of the autorespond program in action reminds me of the chirping that takes place with AMTOR stations, only much slower. It might be described as a "burp" rather than a "chirp!"

programs

Figures 1 and 2 are sample programs reflecting the developments discussed above. Note that these programs, written for the VIC-20 and Commodore 64 computers, were specifically designed to work with the Kantronics interface.

For ease of program development it's necessary to use the RS-232 user port on the VIC-20 and Commodore 64 rather than the joystick port as many commercial software products do. A connection diagram for matching the VIC-20 and Commodore 64 to the Kantronics interface in this manner is shown in fig. 3.

Applying these techniques to other computers will require the writing of individualized programs.

plans for future programs

Our automated propagation study has been very helpful in assessing and continuing to assess equipment requirements and best times for data transmission.

We have found that AMTOR, although somewhat slow, allows us virtually 100 percent accurate transmission as long as we have any signal at all. Knowledgeable hams have walked into the room during AMTOR reception and sworn that no audible tones were present, even as error-free text was being displayed on the screen.

Our experiments have been successful enough that I am now able to run an MSO program for Jon to access and leave messages. A similar operation is planned for his station.

The next step will be a combination of beacon and MSO techniques. We are currently developing software that will leave both stations idle until one or the other station is loaded with a message for transmission. When one or more messages are loaded, the originating station will begin sampling conditions by transmitting a beacon message every five minutes or so. Upon receipt of an acknowledging message from the distant end, the traffic will be transmitted with a check sum. If successfully received, the check sum will be echoed back. The program will allow for several attempts to occur before reverting to the beacon mode. Once the message or messages have been received, both stations will return to monitoring condition.

closing remarks

Why not put that computer to work doing something other than logging contacts? The application described here only begins to suggest the possibilities inherent in using low-cost micros for Amateur Radio application. Perhaps our experience will encourage you to try some of these techniques.

reference

1. Robert B. Rose, K6GKU, "MINIMUF: A Simplified MUF-Prediction Program for Microcomputers," *QST*, December, 1982, page 36.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

a fair shake?

The January 28, 1985 issue of *Business Week* had an interesting article that may be of great import to Radio Amateurs. The "Washington Outlook" column read in part:

Federal Communications Chairman Mark S. Fowler, concerned about the glacial pace of Japanese certification for U. S. communications products, is considering steps to stiffen existing U. S. licensing and regulations requirements. If Fowler lives up to his threat, the rule changes could slow Japanese telecommunications exports to this country to a trickle.

In the same article, Albert Halprin, Chief of FCC's Common Carrier Bureau said, "There is legitimate concern by U. S. manufacturers about whether they are getting a fair shake in other markets."

The article concludes with speculation that Senator John Danforth (R-Missouri) the new Chairman of the Senate Commerce Committee, may reintroduce legislation that would direct the Administration to take similar action against Japanese telecommunications products if U. S. manufacturers are subject to discrimination by Japan.

160 meters revisited

The 160-meter band is in the summer doldrums now, but will spring back to life this fall. More Amateurs are rediscovering this old but interesting band and are determined to operate on it during the coming months. But what antenna can you use on a band whose half-wavelength is about 246 feet at 1.9 MHz — and you live on a small city or suburban lot? Big antennas are great if you live in the country on plenty of acreage, but most hams aren't so lucky.

Paul, W6PYK, wrestled with this problem. His space was limited, and he didn't want to dig up the whole yard to bury a mess of radial wires. He started experimenting in 1983, when he lived in Kentucky, and continued his tests when he later moved to California.

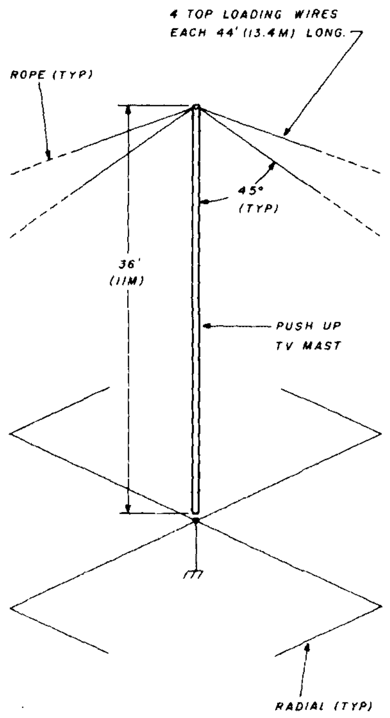


fig. 1. The W6PYK vertical antenna for 160 meters. Antenna is top-loaded TV mast. Base radials slope upward with ends 5 feet (1.5 meters) above ground level.

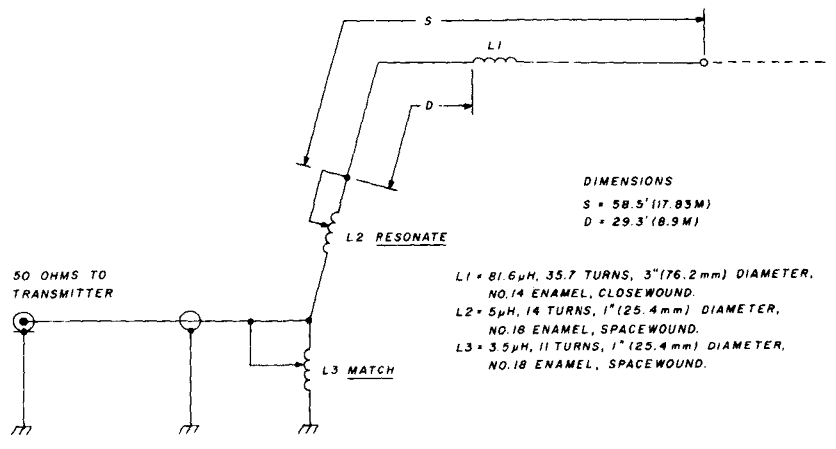


fig. 2. A compact Marconi-type antenna for the 160-meter band. Best results will be obtained with maximum amount of antenna wire in the vertical plane. Height of horizontal portion should be at least 25 feet (7.6 meters).

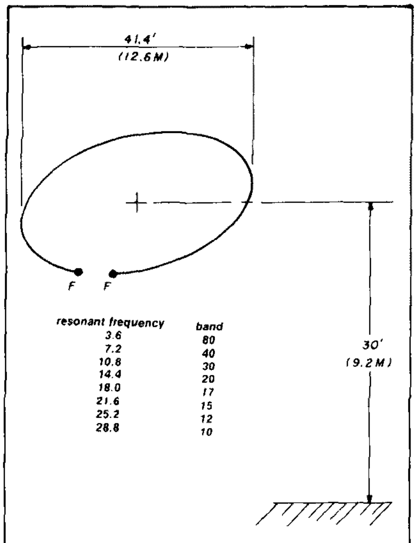


fig. 3. Horizontal loop provides all-band coverage from 80 through 10 meters, including new WARC bands. (Antenna tuner is shown in *fig. 4*.)

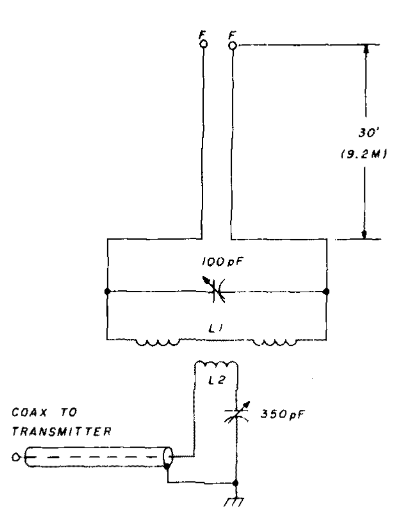


fig. 4. Link-coupled matching network. L1 is a 2-inch diameter coil consisting of two windings of 17 turns per inch. Coil L2 is 14 turns with same diameter. The whole assembly is made from a section of B&W 3026 miniductor stock. Coils are tapped to achieve resonance. (See *The ARRL Antenna Book*, 14th edition, pages 4.5-4.6 for construction details.)

Paul's basic antenna is a 36-foot (11 meter) high push-up TV mast. The theoretical feedpoint resistance of this antenna, working against a perfect ground is 9 ohms at 160 meters. In order to achieve this figure, he used top-loading, as shown in **fig. 1**. The four top loading wires served as guys for the mast.

As a starter, he used a single 8-foot (2.44-meter) long ground rod at the base of the vertical. The feedpoint resistance was about 55 ohms, a good match for a coax line, but a *highly* inefficient setup, as most of the transmitter power was lost in ground resistance. (The ground resistance is the difference between the theoretical feedpoint resistance for the given height and the observed value.) Antenna efficiency, then, is 9/55 = 16 percent.

Paul next disconnected his ground rods and added four base radial wires, close to the ground at the antenna base and about 5 feet (1.5 meters) above ground at the ends. It was necessary to bend the wires into a Z-shape to fit them on the property. The radials were each a quarter-wavelength long.

The feedpoint resistance was now about 13 ohms — a big improvement. Efficiency had risen to 69 percent.

This antenna obviously works. Paul has contacted three continents and many states running only 100 watts. He's now experimenting with a 65-foot (19.8 meter) tower in the same configuration.

a compact 160-meter Marconi antenna

When you have no room for a big antenna, small is best. You may not be Number One on the frequency with a small antenna, but you're on the air and can have plenty of fun. Here's a design for a coil-loaded Marconi antenna for the "top band" (**fig. 2**). The antenna is self-resonant at 2 MHz with the center loading coil L1. Series coil L2 at the feedpoint drops the resonant frequency as low as 1.8 MHz. And shunt coil L3 provides a match to a 50-ohm feedpoint.

An antenna can't be much simpler than this one. Its overall length is only 58.5 feet (17.8 meters). The antenna is bent into an "L" shape, with the horizontal portion 25 to 30 feet (7.6 to 9.2 meters) above ground. It has been used with success with the continuous metal plumbing system of the residence acting as a ground.

Adjustment is simple. Coil L2 tunes the antenna to resonance and coil L3 provides the correct impedance transformation to a 50-ohm feedpoint. The adjustments are slightly interdependent, but can be quickly accomplished with the aid of an SWR meter. Antenna operating bandwidth between the 2:1 SWR points on the feedline is about 50 kHz.

an "all-band" horizontal loop antenna

The virtues of voltage feeding an antenna have not been fully appreciated by the Amateur fraternity. When the voltage fed antenna is bent into a loop, a very interesting antenna results (**fig. 3**). This illustration shows a horizontal loop antenna about 130 feet (39.6 meters) in circumference. The harmonic resonant frequencies listed in the chart (see **fig. 3**) show that the loop provides resonance at, or near, the Amateur bands between 80 and 10 meters.

If desired, the loop can be made a bit smaller, with a portion of it being the feedline.

A simple antenna tuner is required to match the loop to a 50-ohm line (**fig. 4**). Resonance and coupling controls are adjusted in order to provide the lowest SWR at the transmitter.

Experimenters using this antenna will find that the resonant points are very broad at the higher end of the spectrum and that the resonant frequency of the loop can be "pulled" to almost any spot in the HF spectrum.

A loop twice this size, with a circumference of 260 feet, or 80 meters, exhibits twice the number of resonant frequencies and operates well at any frequency between 1.8 and 30 MHz with the proper antenna tuner. In all instances, the feedpoint of either antenna is at a high impedance.

The antenna need not be a perfect circle; it can be a many-sided polygon which encloses as much area as possible.

A simple and convenient feeder can be made up of a short length (30 feet, or 9 meters) of 300-ohm "ladder line" or perforated twin lead.

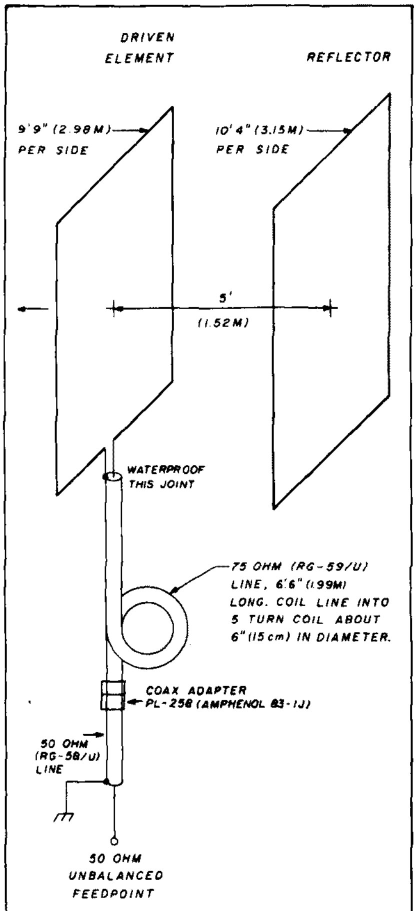


fig. 5. Simple 2-element Cubical Quad antenna for 24.9 MHz band.

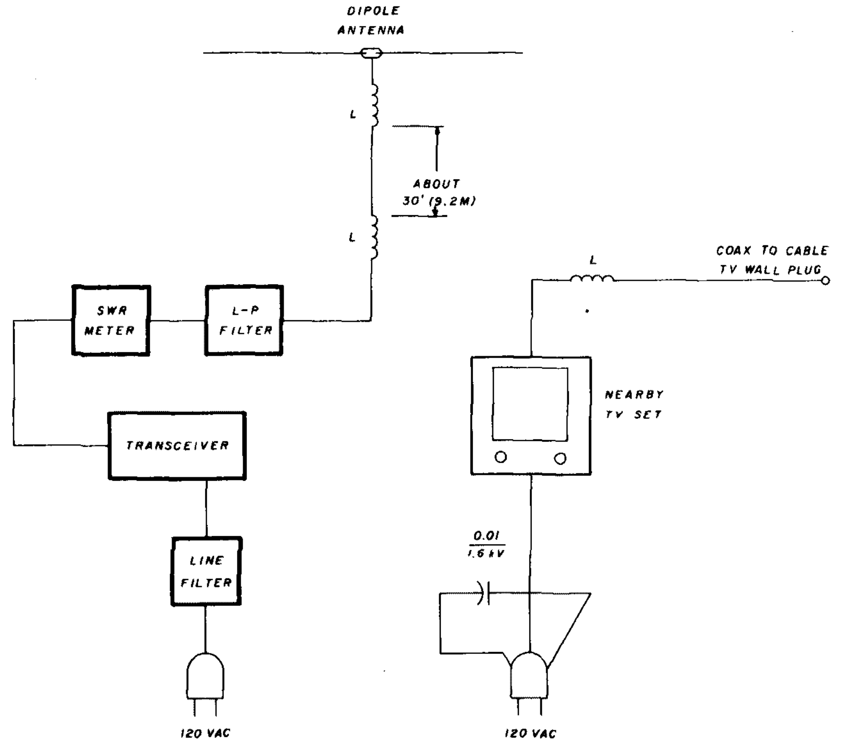


fig. 6. Line filters, plus winding coax lines into RF choke coils (L) clean up residual TVI after low-pass filter is placed in line with the transceiver.

a two-element quad for the 24.9-MHz band

With the opening of the 24-MHz band, there is considerable interest in a simple beam antenna that can be easily constructed and will provide good gain. The old favorite, the Cubical Quad, provides an inexpensive solution (**fig. 5**). The Quad provides a gain of about 7 dB over a dipole with a front-to-back ratio of approximately 15 dB. A quarter-wavelength transformer made of a section of 75-ohm line provides a good match to the Quad. The line is coiled up into an RF choke to reduce currents flowing on the outside of the outer shield of the line.

TVI revisited

How to clean up a bad case of TVI at a resort condominium? I had that problem last spring while on vacation. The rental condo had cable, but that didn't prevent the TV from going black and making funny noises when I fired up my transceiver on 20-meter SSB. Being prepared for such an eventuality, it took less than an hour to set things right. Here's what I did:

1. I placed a line filter — a three-section J. W. Miller, No. C-508-L — on the transceiver. I didn't have a second line filter for the TV receiver, so I made a simple one consisting of a line plug with a 0.01 μF, 1.6 kV disc ceramic capacitor wired across the prongs. I plugged this into the same outlet that fed the TV.

2. I then placed a low-pass filter in the coax lead from the transceiver to the antenna (a dipole). The filter was placed after the SWR meter, since the diodes in the meter can often generate TVI when it is not otherwise present.

3. The final step was to wind the coax (RG58C/U) into an RF choke at the point at which it joined the antenna. I made a five-turn coil, about six inches (15 cm) in diameter, held in position with electrical tape. A similar coil was made in the transmission line at the station end, just after the low-pass filter.

The installation is illustrated in **fig. 6**. It did the job! The TV was clean on all channels. If the TV had been on an antenna instead of on cable, a high-pass filter at the input terminals of the TV would probably have been necessary.

Coiling the coax line from transmitter to antenna into simple RF chokes was an important part of the solution. Without the coils, interference was noticeable on channels 2 and 4. When the chokes were in the line, the interference disappeared.

When the vacation was over, it was but the work of a moment to drop the dipole and remove the filter capacitor from the TV power plug. The two RF choke coils were left permanently in the dipole feedline and the Miller line filter was packed away with the transceiver for use on the next vacation trip.

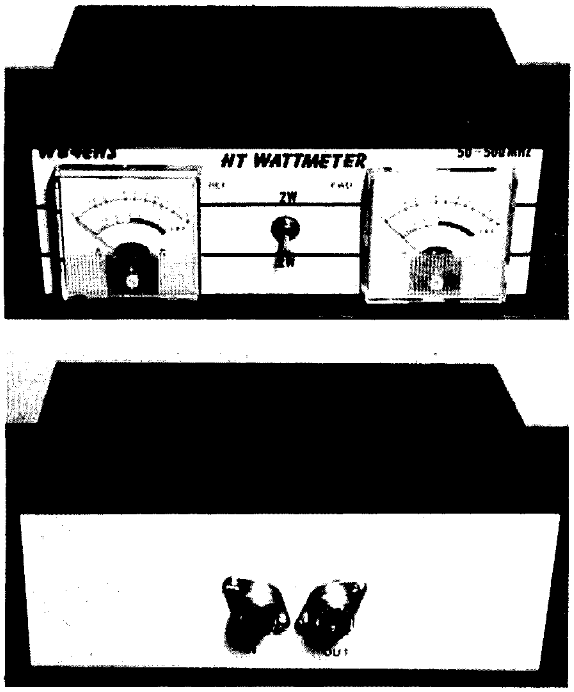
ham radio

the weekender

a 50-500 MHz dual wattmeter

Having been interested in the 420-450 MHz band for some time, I finally succumbed to the UHF "bug" and bought a commercial FM HT to get on this band. Shortly thereafter I decided to build some sort of directional antenna system to make my HT more versatile. But a major stumbling block immediately arose: I had no SWR or power measuring capabilities for these frequencies. Because all the designs I could find called for assembly skills I do not possess, I decided to design my own device (**fig. 1**).

The heart of this circuit is the directional couplers (Mini-Circuits Labs No. PDC-10-1). These are 11.5 dB couplers (the sampled port is −11.5 dB down from the line) encapsulated in miniature metal cans. The coupling ratio is flat to within ±0.6 dB from 500 kHz to 500 MHz, and maximum power on the throughline is 3 watts from 5 to 500 MHz (1.5 watts below that). Thus this circuit will measure low power from 5.0 to 500 MHz directly.



Front and rear views of the completed wattmeter. Lettering is rub-on type; striping is 1/16-inch PC layout tape.

item	description
C1,C2	0.01 μF ceramic capacitors, 15 volt
CR1,CR2	HP-2800 or equivalent Schottky diodes
M1,M2	0-100 μA meters
R1-R4	50 kilohm trimmer potentiometers
U1,U2	Mini-Circuits PDC-10-1 directional couplers

Sources:
HP-2800: Megahertz Electronics, 2111 West Camelback Road, Phoenix, Arizona 85105
PDC-10-1: Mini-Circuits, P.O. Box 166, Brooklyn, New York 11235. Price: approximately $11.50 each in 5 to 45 quantity. (Singles about double that price.) Club or individual orders accepted.
0-100 μA meters — calibrated: Fair Radio Sales, 1016 East Eureka, Box 1105, Lima, Ohio 45802.

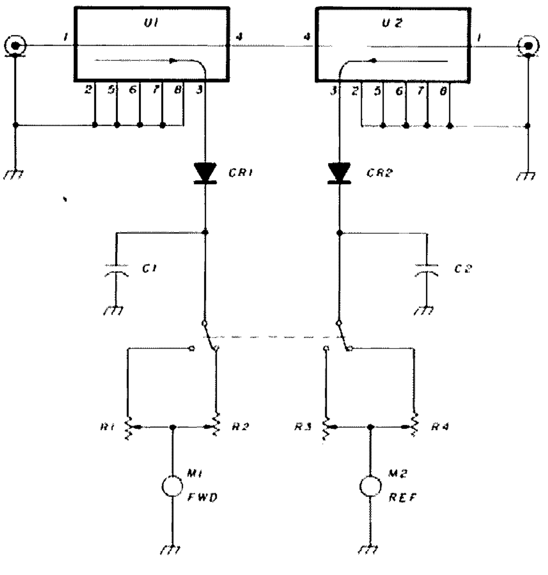


fig. 1. Dual wattmeter schematic.

The remainder of the circuit is a typical RF voltmeter. The HP-2800 microwave diodes rectify the sampled RF and charge the 0.01 μF capacitors. A DPDT switch selects trim pots for the ranges of 0.2 and 2.0 watts full scale. This set of ranges was chosen because meters calibrated from 0 to 20 watts, with a 100 μA movement, were readily available. (To make this a peak-reading wattmeter for SSB, replace the 0.01 capacitors with 6.8 or 10 μF electrolytics.)

construction tips

I strongly recommend the use of good quality double-sided fiberglass PC board, with one side (the bottom) etched and the other containing non-grounded holes countersunk to prevent shorts. All components are mounted on the ground-plane side. In the vicinity of the throughline, which will be carrying up to 450 MHz energy, I put several "Z" wire jumpers between top and bottom ground planes to prevent ground problems. After using the meter for several days, I added a shield of brass sheet (shim stock) over the RF throughline portion of the card. I can't really say I noticed much of a difference as a result.

By Bob Lombardi, WB4EHS, 1874 Palmer Drive, Melbourne, Florida 32935

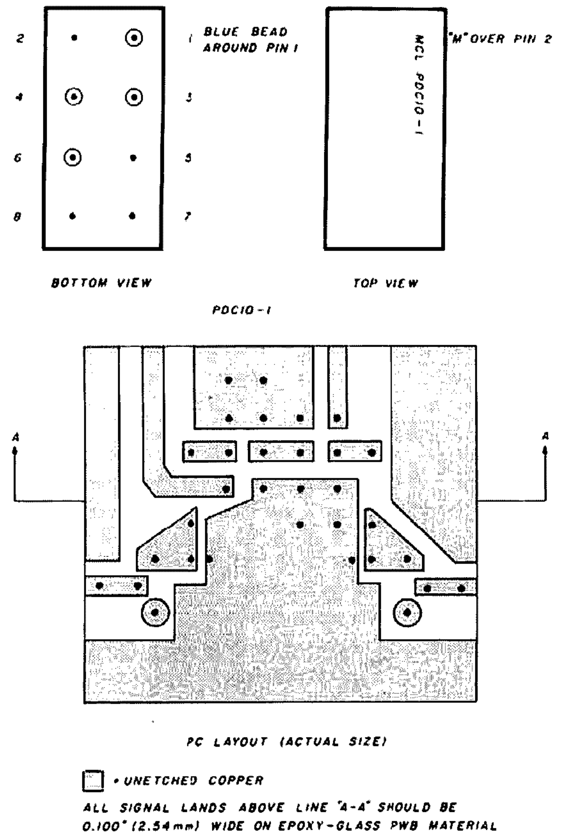


fig. 2. PC board layout (actual size).

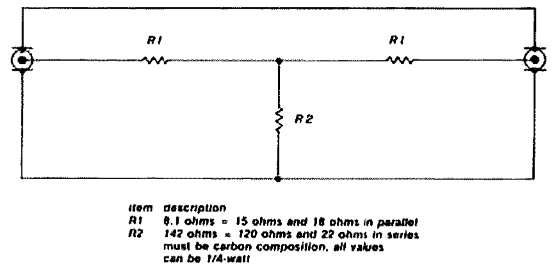


fig. 3. 3dB pad is a useful addition to meter.

It will be noted from the PC layout (**fig. 2**) that the directional couplers are mounted offset from parallel and case-to-case. The fact that one coupler is used for reverse voltage sensing dictates that they must be mounted with their part-markings in opposite directions. This minimizes the path that the RF must take through the meter. If you plan to use the meter only on 2 meters, this extra care in layout and shielding could probably be omitted.

You may notice that I've used SO-239 connectors instead of BNCs. Up to about 500 MHz, connector choice is a matter of just that — choice. A well-installed (i.e., short grounds with good shielding all around) RCA phono plug is as good as a BNC plug. Because I already had the connectors on hand, and am already good at installing them, I stayed with SO-239s. For higher power work, or serious weak-signal work on 432, I would probably switch to N connectors.

calibration and use

Since the meter response can vary with frequency, it's best to calibrate it on the band in which you are most interested. Terminate the output with a 50-ohm non-inductive resistor and calibrate the forward position to whatever your standard is known to be. I used a Bird Termaline wattmeter as standard at 446 MHz. Interchange the input and termination and do the same on the reverse meter. Do this on both ranges.

The calibration of this type of meter depends on line impedance. With purely resistive loads.

$$P = I^2R \text{ or } P = \frac{E^2}{R}$$

When R is not the 50 ohms with which we calibrated, the accuracy falls off. For any power measurements we make with the meter, it should always be terminated in 50 ohms resistive.

For tweaking antennas, a familiar equation is:

$$VSWR = \frac{1 + \sqrt{\frac{P_{REF}}{P_{FWD}}}}{1 - \sqrt{\frac{P_{REF}}{P_{FWD}}}}$$

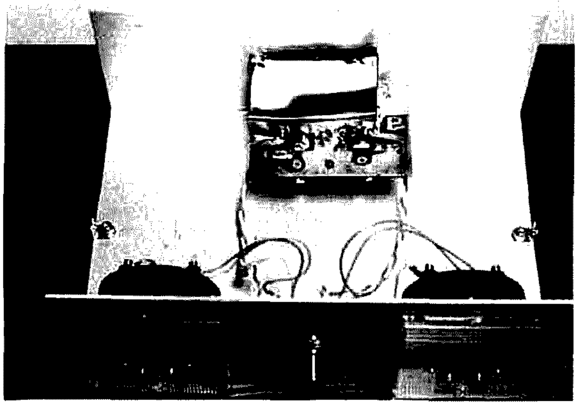
Charts are available in the literature for fast determination of VSWR for a given P_{REF} and P_{FWD}. This can be also done in seconds on a simple calculator. In most cases, however, all that's necessary is to observe the forward increase and reverse decrease in readings while working on the antenna (nulling VSWR).

In UHF work, it's important to remember that the SWR at the antenna will always be worse than the SWR at the meter unless the antenna and meter are very close — for example, if the antenna is mounted on the back of the meter box. For example, suppose you measure 2 watts forward and 0.05 watts reverse on a section of coax with 3 dB loss. Using these values, the equation gives us $SWR = 1.4{:}1$, which is probably reasonable. Taking the 3 dB loss into account, the forward power at the antenna is 1 watt, and the reflected power is 0.1 watt, (the 3dB works both ways). This gives an SWR of 1.9:1, which may or may not be acceptable.

For antenna testing, measure as close to the antenna as possible, and know (or better yet, minimize) losses.

other measurements

This circuit is useful for measuring powers above 2 watts if they are reduced to the 0-2 watt range before



Because my original version was only 2 watts full-scale, the PC layout shown in *fig. 2* is different. It's convenient to mount the pots on the switch; if you opt to do this, change the PC layout to accommodate the two sets of trimmers.

being applied to the meter. **Figure 3** shows a 3-dB attenuator for reducing 4 watts to 2 watts; this should handle just about all of the commonly used 2-meter HTs. It's best to calibrate your attenuator on your wattmeter by measuring power level of a known $\leq$ 2 watt source both with and without it in line. Mark the power factor (about two times) on the pad and multiply power measurements by that factor whenever using it. Of course, it's safest to take the first readings of any new source with the pad in line. It can then be removed if the source is less than 2 watts. Other types of couplers can be used to measure higher power levels: for example, a 10-dB coupler can be used to measure up to 20 watts, and a 20-dB coupler can be used for up to 200 watts.

conclusion

In-line wattmeters should not be used in VHF or UHF weak signal work because the losses are not tolerable. This unit works well in its intended applications — low power measurements and antenna tweaking. Likewise, with the proper choice of attenuators and couplers, it's useful for measuring other power levels in the 5 to 500 MHz range.

Mini-Circuits claims an insertion loss of 0.85 dB per coupler, or 1.7 dB for the meter. Input VSWR should be 1.2:1. I measured VSWR for my unit during calibration and found it to be about 1.3:1, within reason when the connectors and adaptors used are figured in. I wasn't able to measure insertion loss because of the plethora of cables and adaptors required.

This is an unusual circuit, the only one I'm aware of that extends alternative techniques to UHF in Amateur applications. All other Amateur circuits I could find required brass pipe and other hardware. There are plenty of 2-meter, 220 MHz, and 440 MHz HTs out there, with powers in the range of this instrument; I hope their owners find this project useful.

ham radio

short circuits

July 1985

carrier suppression

In the April ham note (page 78), "Improved Carrier Suppression for the MC1496," *power* — not signal, as printed — is applied to the bifilar windings through a series choke arrangement.

July 1985

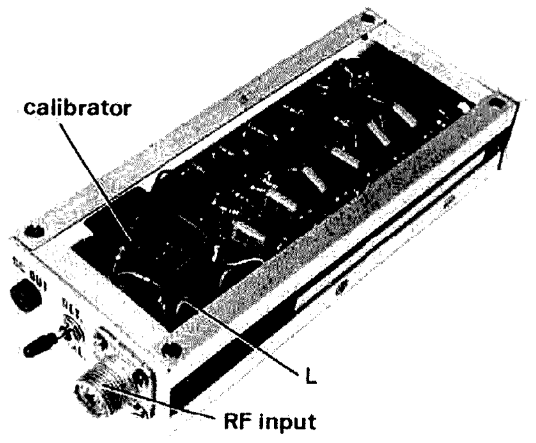
feeding phased arrays

The caption for **fig. 1B** of KB8I's article, "Feeding Phased Arrays: An Alternative Method" (May, 1985, page 59) should be revised to read as follows:

fig. 1B. Though high SWR exists on the main feeder, the matchbox now located in the shack can be switched over to and used with other antenna systems.

wideband logarithmic detector

Only six transistors provide a 60 dB log response



Logarithmic detector consists of six amplifier-detector stages in cascade. Detected RF input signal is processed to achieve a logarithmic function over a 60-dB range.

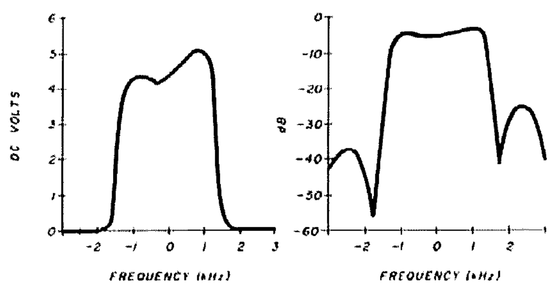


fig. 1. Two curves of the same bandpass filter characteristics. Left: as displayed when using a diode detector (voltage-linear scale and diode threshold of 0.4 volt). Right: as displayed when using a logarithmic detector.

The question of why a logarithmic detector would be desirable for filter sweep alignment may be best answered by considering the virtues of the decibel scale. With a logarithmic detector an oscilloscope display can show parts of the filter characteristic that an ordinary detector would probably ignore. The display may even reveal an entirely different and more realistic picture of the test results (**fig. 1**).

Several methods are available for building a logarithmic detector. For example, one could consider using the nonlinear behavior of a diode (very much temperature dependent); a string of clamping diodes for correcting the DC detector output (amplitude-range limitations); using an LM3089[1] (which, unfortunately, decides to stop working properly below 10 MHz); or making a succession of amplifiers and detectors with the detector outputs connected in parallel.

I chose the last method. What may be unusual about this approach is the use of amplifiers that are not tuned to one particular frequency, as opposed to what is customary in applications such as spectrum analyzers or field strength recorders. These are wideband amplifiers, accepting signals with frequencies between 50 kHz and 14 MHz. It is not only the absence of coils that makes this little test box so simple. Only six transistors are necessary to realize a reasonably accurate logarithmic response over a range of almost 60 dB. **Figure 2** shows the general arrangement of the logarithmic detector. Final test results are depicted in **table 1**.

the principle

Starting off by experimenting with a common diode detector, I found that by adding a reversed-polarized diode, counteracting the curvature of the detector characteristic, a very reasonably logarithmic curve portion can be extracted, extending over a range of almost 10 dB (**fig. 3**). Six amplifier-detector stages are connected in cascade. As each stage increases the RF signal amplitude by 10 dB, each detector processes

By Hans Evers, PA0CX/DJ0SA, Am Stockberg 15, D5165 Hürtgenwald, West Germany

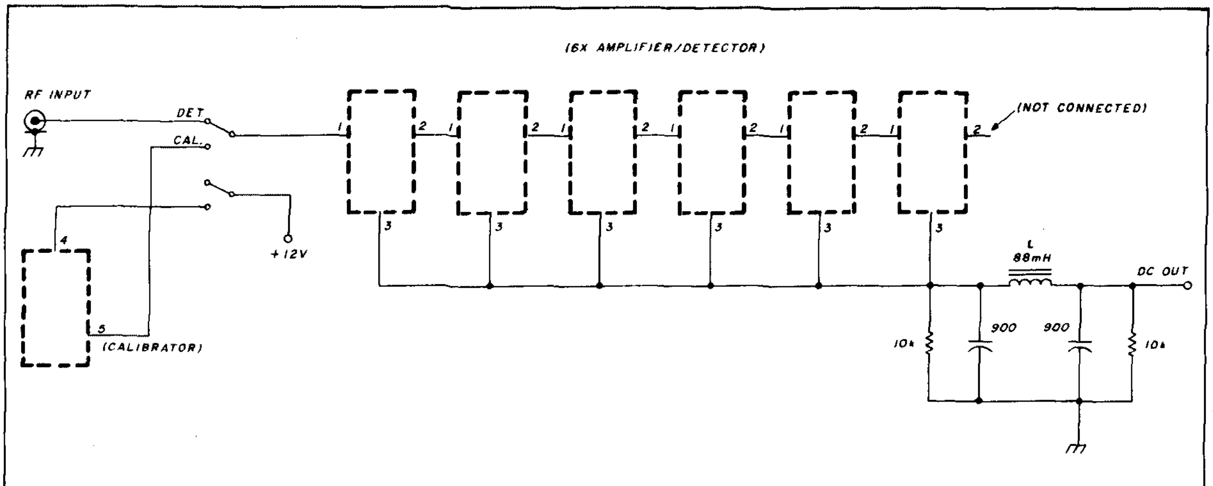


fig. 2. Schematic diagram of logarithmic detector. Circuit provides a 60-dB dynamic range. (See figs. 4 and 6 for details of amplifier/detector stages and calibrator unit.)

the signal with mainly the logarithmic portion of its characteristic. The detector outputs are combined, partly overlapping, compensating for the less desirable portions of their characteristics. This translates to a 60 dB range logarithmic conversion of an RF inputted signal. The reason for the 60-dB range is that it looked about right for Amateur Radio use. It covers the 9 S-units, plus an extra 6 dB. In practice, this represents sufficient dynamic range for determining carrier and unwanted sideband suppression, filter shape factors, and for detecting side lobes of crystal filters.

transient response

CR1 and C1 form a peak detector (**fig. 3**). With R1 given, C1 must be adequately large for charging to the full signal peak voltage. If it is not, the detected DC voltage is no longer a true function of the RF amplitude. If, on he other hand, the RC time constant is too long, the circuit may not be capable of following the transients — *e.g.*, of those caused by the possibly steep skirts of a filter swept at a high rate.

In this case, the value of 0.001 μF for C1 allows an acceptable compromise between the lowest frequency (50 kHz) at which I decided that the detector should still be usable, and a transient response that should not take more than 0.5 millisecond to be fully displayed. In practice this means that using a (flicker-free) time base of 20 times per second, the amplitude could make a full-range jump in less time than it would take to displace the oscilloscope light spot horizontally by 1 percent. This makes the detector reasonably fast, so that it can even be used as part of a spectrum analyzer or panoramic receiver.

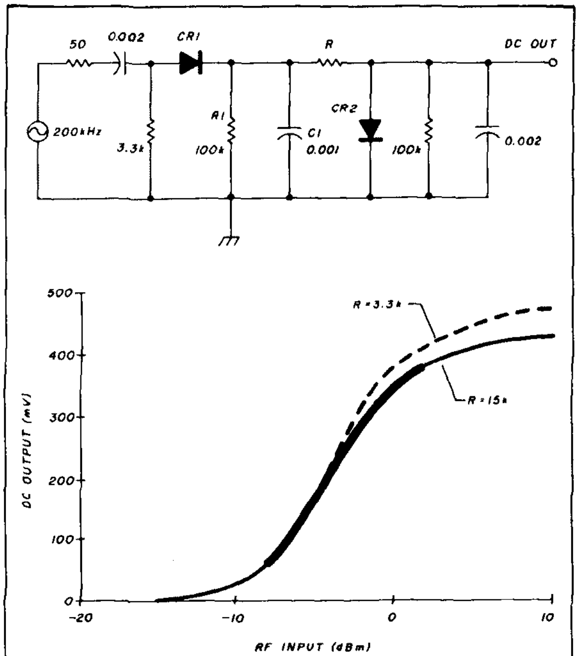


fig. 3. Effect of a clamping diode CR2 on detector characteristic.

amplifier

Each amplifier stage has been designed for a gain of exactly 10 dB over practically the whole range of intermediate frequencies used in Amateur Radio equipment, extending from the old 50-kHz "Q-fiver" to the more modern 9- and 10-MHz transceiver filters. Further requirements were a low output impedance

(about 50 ohms) to avoid loading by the detector with its varying impedance, and a dynamic range with a few dBs to spare before the transistor saturates. This, as well as the bandwidth and stability, has mainly been obtained by applying heavy negative feedback.

With only 100 pF across the emitter resistor of each amplifier, the bandwidth is about 2 MHz at the −0.5 dB points. However, by adding extra capacitance the bandwidth can easily be increased to 14 MHz without deteriorating the flat frequency response. This brings

table 1. Final test results.

frequency range:	50 kHz to 14 MHz (−1 dB)
RF input:	−60 dBm to 0 dBm (50 ohms), or 0.22 mV to 0.22 V (500 ohms)
DC output:	0 to 120 mV (oscilloscope sensitivity of 20 mV/division at 10 dB)
maximum error:	1.5 dB
power supply requirements:	120 volts/100 mA

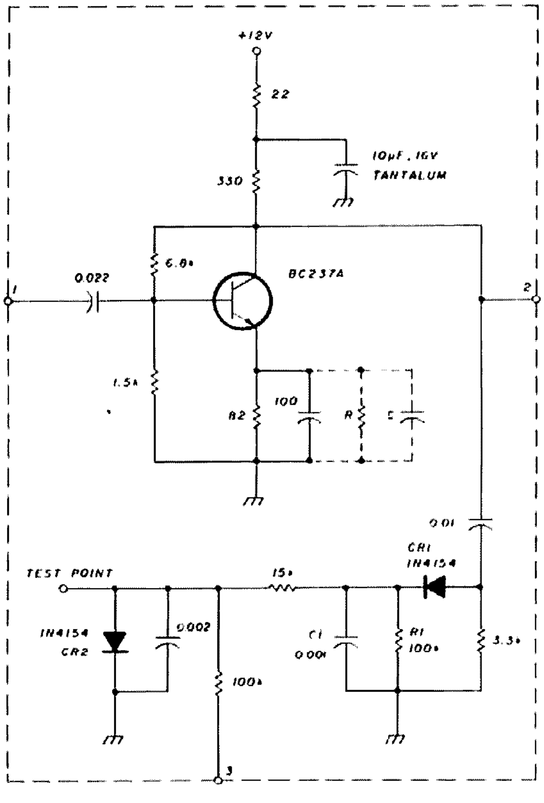
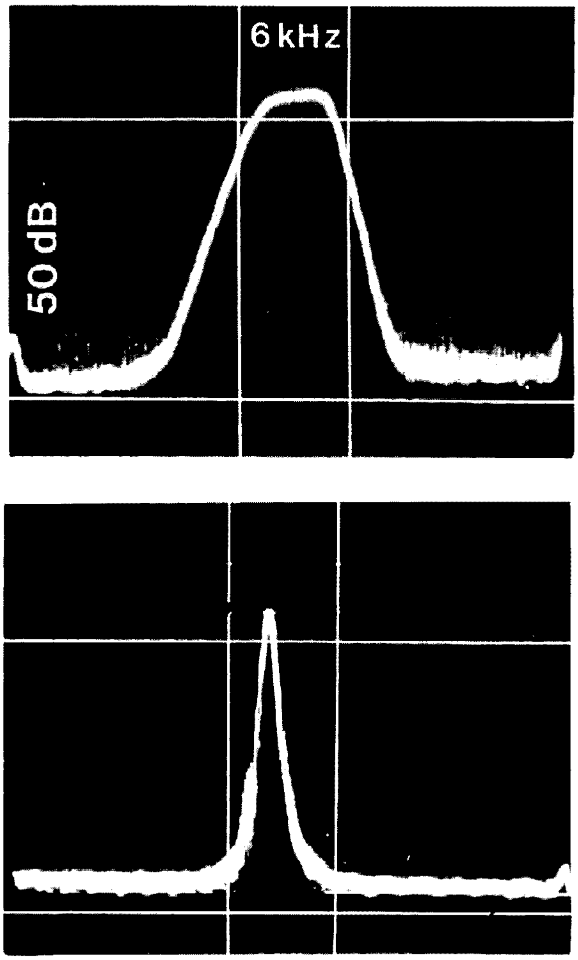


fig. 4. Amplifier/detector stage. Resistor R may be necessary only for correcting the amplifier gain; capacitor C may vary between 50 and 200 pF and has to be selected for optimum bandwidth. (See text.)



Using the logarithmic detector for sweeping an IF amplifier. 352-kHz IF bandpass crystal filter with continuously variable bandwidth in WWII German receiver (Mw.E.C.). Top: at maximum bandwidth (5 kHz); bottom: at minimum bandwidth (130 Hz) (sweep rate 5 Hz).

the detector perfectly in line with the compact IF sweep generator published last month in *ham radio*.[2]

Only readily available components were used. The BC237A is a rather popular audio transistor (at least here in Europe), and the only reason why the 1N4154 was chosen was that, at the time, it was the least expensive diode available in the local parts shop. A perhaps more common 1N914 or 1N4148 would probably work just as well. Normal 5 percent resistors were used (note that 5 percent resistance means 0.5 dB tolerance), yet only a slight correction was necessary to get the voltage gain of each stage at exactly 3.16 times (10dB).

The input impedance of the logarithmic detector is about 500 ohms. In case a 50-ohm input impedance is desired, merely connect a 120-ohm and a 100-ohm resistor in parallel with the RF input plug.

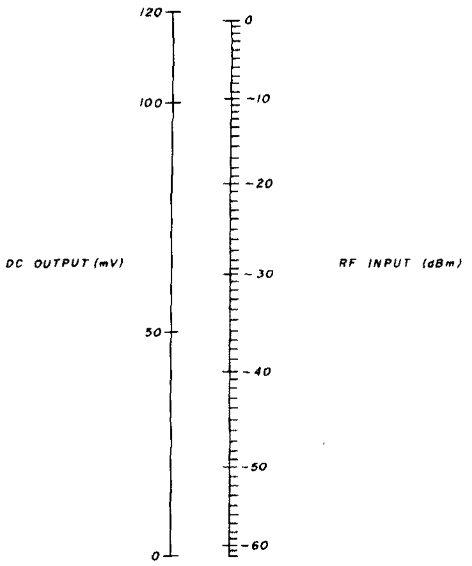


fig. 5. DC output voltage as function of RF input level.

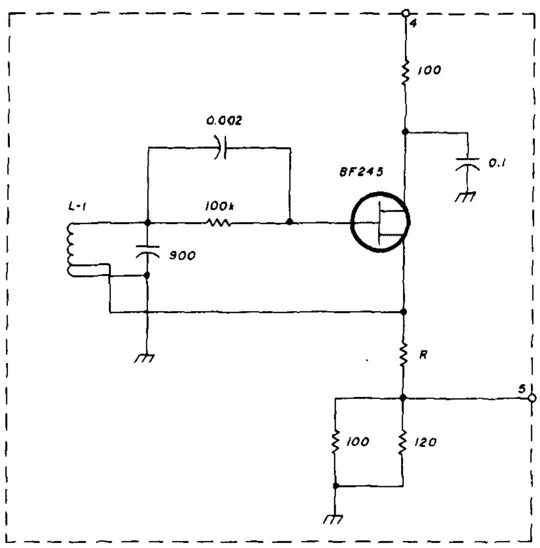


fig. 6. Calibrator. R is approximately 50 ohms, to be selected to obtain a calibrator output of 0.22 volts RMS with 500 ohms load. L1 has 80 + 6 turns on Amidon T-3-50.

The ripple residue on the DC output voltage is imperceptibly small. Only for RF input signals at frequencies lower than a few hundred kHz, the picture lines on the scope tend to become a bit woolly. The LP filter, consisting of an 88-mH coil and two 900-pF capacitors, also removes that last remainder of RF. Its influence begins around 20 kHz, far beyond the point at which it could limit the transient response. At 28 kHz the ripple is already suppressed by 20 dB. In case the detector is never used below, say 300 kHz, the coil and two capacitors could be omitted, leaving the detector without any coil at all.

trimming procedure

This consists of applying an RF signal of about 1 volt to the input, thereby saturating all detector stages. The signal frequency should be somewhere between 100 kHz and 1 MHz. Using a high-impedance voltmeter, measure the DC voltage at each detector test point and see that all are equal **(fig. 4)**. If not, increase a possibly low voltage be selecting resistor R in parallel with the 82-ohm emitter resistor of the corresponding amplifier.

To achieve the full 14-MHz bandwidth, apply a 10-MHz signal, saturating all detectors again. Increase the possibly low test-point voltages *this time by* paralleling an extra capacitor across the emitter resistor of the deficient amplifier. **Figure 5** shows how the end result looks after applying the above procedure. The regularity of the dB scale divisions is well acceptable, as shown, with only a slight compression at the top and bottom end. This effect is difficult to avoid; it is caused by the (only partly compensated) first and last detectors lacking the correcting overlap of a neighbor.

calibrator

The logarithmic detector works satisfactorily, is simple and inexpensive to construct, and requires no sophisticated parts or test equipment. If this seems to be almost too good to be true and you're wondering, "where's the catch?" you're right. There is a *"catch," and it's in the* CR2 diodes.

The textbooks state that the voltage drop across a silicon diode junction decreases by about 2 to 2.5 millivolts for every degree centigrade rise in temperature. Here there are up to six of them, *all adding up.* Although the effect is eventually reduced by voltage division, it still works out to a few millivolts on the end result.

This explains why, on a really chilly day, one may find that the whole dB scale has stretched somewhat and the vertical sensitivity of the scope has to be reduced by 5 or 10 mV to restore the calibration of 10 dB per division. This minor inconvenience (incidently, it is the only "warming-up" effect noticeable) does not justify spoiling the simplicity of the design by the addition of extra compensation circuitry.

Nevertheless, those who would feel more bothered by possible level inaccuracy than by the discomfort of an under-cooled Amateur station, may wish to

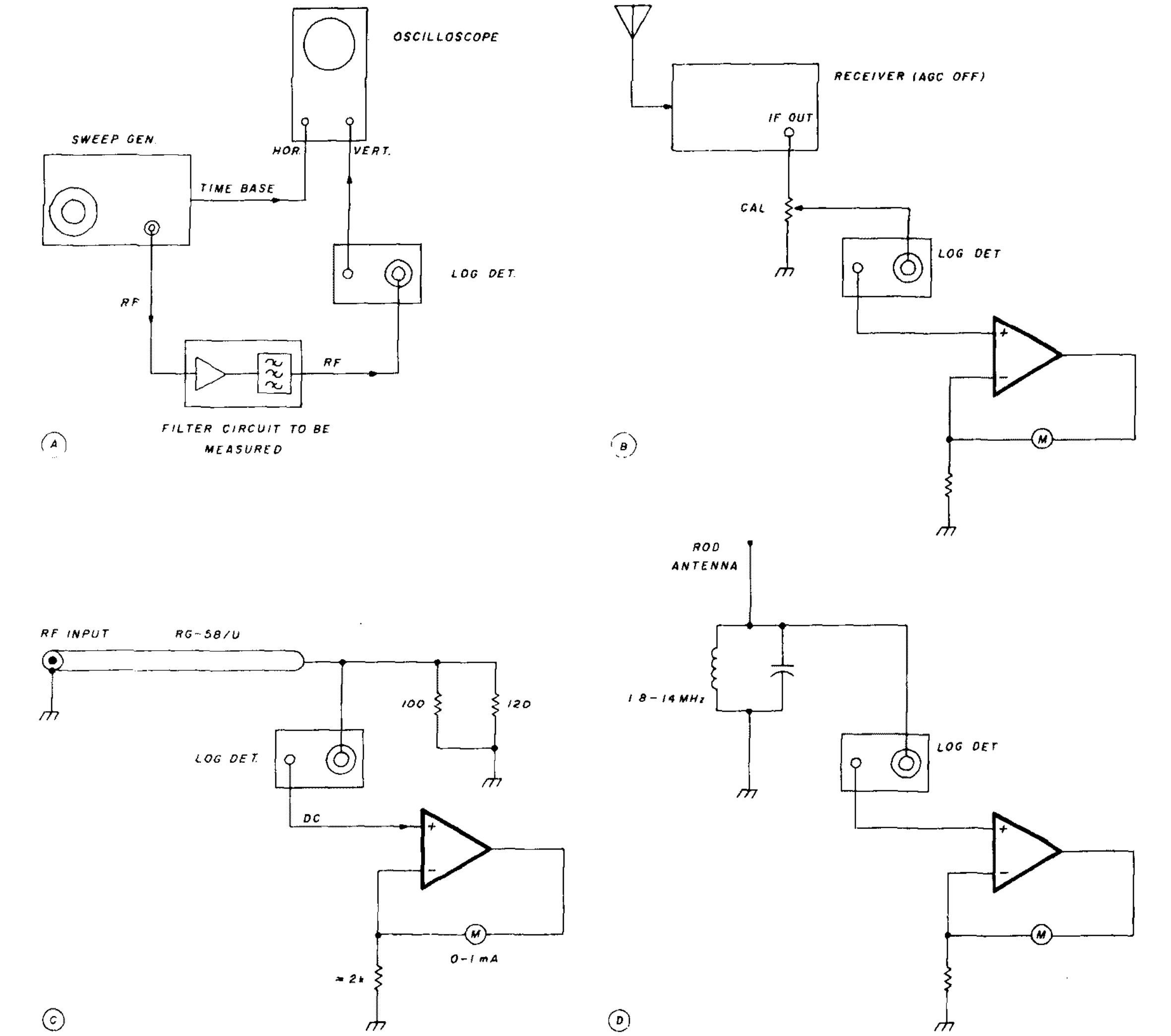


fig. 7. Some applications for a logarithmic detector: (A) Measuring frequency response of an RF filter; (B) Reliable "S" meter, S1 to S9 + 6dB, (C) Single range RF microwattmeter, 2 nanowatts (at 3 percent FSD) to 1 milliwatt (at 100 percent FSD); (D) 60 dB field-strength indicator.

incorporate a built-in 0-dBm calibrator (**fig. 6**). Once the 0-dBm level is set at the correct height of six divisions on the scope screen, the other decades fall into place by themselves. The device consists of a sinewave oscillator that provides a 500 kHz constant amplitude source of energy. Calibration is not difficult because the oscilloscope itself could be used for the initial setting of the 0-dBm calibrator output voltage. Just remember that 0-dBm into 50 ohms corresponds to 0.62 volt peak-to-peak (0.22 volt RMS).

applications

The application of the logarithmic detector is not limited to sweeping filters only. As some examples in **fig. 7** show, it could be the backbone for an RF microwatt meter with a linear dB scale, or the development of a truly reliable S-meter. Adding a simple tuned circuit results in a deluxe field-strength meter for the HF bands down to 20 meters.

references

1. R. Ferranti, WA6NCX/1, "Design Notes on a Panoramic Adapter," *ham radio*, February, 1983, page 26.
2. Hans Evers, PA0CX, "Compact IF Sweep Generator," *ham radio*, June, page 35.

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

propagation update

When it comes to discovering new propagation modes and extending DX records, the VHF/UHF/SHF frequencies represent Amateur Radio's latest frontier. Judging from correspondence and on-the-air discussions, there's a pioneering spirit — and a great curiosity about the unknown — among hams active on these bands.

Because reader response to last July's column[1] on VHF/UHF propagation was so encouraging, I've dedicated this month's column to expanding and updating the material presented in that issue.

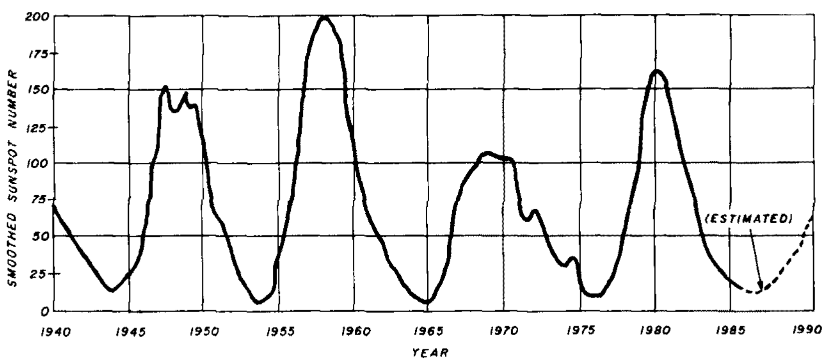


fig. 1. This graph shows the variation in the smoothed sunspot numbers over the last few solar cycles.

VHF/UHF/SHF frequency bands

It wasn't too many years ago that all frequencies above 40 GHz were open to Radio Amateurs. However, as research in millimeter waves increased, commercial and government interests forced subdivision of the frequencies between 40 and 300 GHz. From our point of view, this measure created new bands to explore as separate entities.

Table 1 shows all the major VHF/UHF/SHF bands available to Radio Amateurs. WARC also subdivided these frequency assignments by IARU regions. Generally speaking, Region 1 includes most of Africa, Europe, and the Soviet Union. North and South America as well as Hawaii are in Region 2, and the Southern portions of Asia and the Oceania nations are in Region 3.

In recent years there has been a tendency to designate all bands in meters rather than by frequency in MHz. I've therefore listed the metric band designations next to each frequency assignment on the table. Note that in some cases a band may or may not be available in some regions. For instance, although UK Amateurs have a 4-meter band, they have no frequencies available between 2 meters and 70 cm. Region 2, on the other hand, has a 6-meter band and a 135-cm band. Where known, I've listed differences in frequency assignments. Normally speaking, the differences are not significant unless you are operating outside your region — such as in EME!

VHF/UHF/SHF DX records

By the time last July's *ham radio* appeared, many of the VHF/UHF/SHF DX records listed in my column, prepared two months earlier, had been broken. To me, DX records — especially in the world above 50 MHz — are a major driving force in the progress of communications technology. At a glance, they reveal not only the development of the state of the art, but also suggest the possibilities and challenges available.

The VHF/UHF/SHF DX records claimed worldwide and given in last year's July column have been, with the exception of those claimed for EME, updated and listed in **table 2. Table 3** shows EME record claims. In recent years many of the records once held in Region 2 have shifted to other regions because of special propagation modes such as transequatorial (TE) or tropo ducting favoring these other regions. This has tended to discourage DXing in those areas in which only more conventional propagation modes such as aurora and meteor scatter are available. So I'll try a new approach (one already used, by recordkeepers in Region 1), listing, in **table 4**, only North American DX claims. This table lists DX records by propagation modes for all VHF/UHF/SHF bands on which Amateur comunications records have been claimed.

I hope that this type of listing will inspire increased effort and exploration of the various propagation modes, especially here in North America. The data listed in **table 4** was difficult to

table 1. Major VHF/UHF/SHF worldwide Amateur Radio frequency assignments.

band	frequency range	notes
6 meters	50-54 MHz	CW only between 50.0-50.1 in USA. Only a few assignments in Region 1.
4 meters	70.025-70.5 MHz	Primarily United Kingdom.
2 meters	144-148 MHz	CW only between 144.0-144.1. Except in Region 2, most other countries have only 144-146 MHz.
135 cm	220-225 MHz	Region 2 only.
70 cm	420-450 MHz	Region 2, Canada only 430-450. Most of the rest of world has only 430-440 MHz.
33 cm	902-928 MHz	Not yet available in USA except for those with FCC experimental licenses. Canada has same but on A3/F3 only.
23 cm	1215-1300 MHz	1215-1240 still available in Canada, but recently withdrawn in USA. Some countries in Region 1 do not have the full assignment. Others have power or EIRP restrictions.
13 cm	2300-2450 MHz	2310-2390 was removed for USA on November 6, 1984. Many Europeans cannot operate below 2320. Japan has only 2400-2450.
9 cm	3300-3500 MHz	Some area restrictions apply. UK has 3400-3475.
6 cm	5650-5925 MHz	Some area restrictions apply. UK has 5650-5850.
3 cm	10.0-10.5 GHz	
12 mm	24-24.25 GHz	24-24.05 in West Germany
6 mm	47-50 GHz	47-47.2 in West Germany, 48-50 in USA
4 mm	71-76 GHz	75.5-76 in West Germany
2 mm	142-170 GHz	165-170 GHz in USA, 142-144 GHz in West Germany.
12 μm	240-250 GHz	248-250 in West Germany.
10 μm	300 GHz and above	No restrictions in USA

table 2. Claimed VHF/UHF/SHF terrestrial DX records (worldwide). EME records are shown in *table 3*.

band	record holders	date	mode	DX miles	DX (km)
6 meters	(see Note 1)				
4 meters	GW4ASR/P-5B4CY	June 7, 1981	E_s	2153	(3465)
2 meters	I4EAT-ZS3B	March 30, 1979	TE	4884	(7860)
135 cm	KP4EOR-LU7DJZ	March 9, 1983	TE	3670	(5906)
70 cm	KD6R-KH6IAA/P	July 28, 1980	ducting	2550	(4103)
23 cm	KH6HME-N6CA	June 24, 1984	ducting	2472	(3977)
13 cm	VK5QR-VK6WG/P	February 17, 1978	ducting	1170	(1883)
9 cm	G3LQR-SM6HYG	July 11, 1983	ducting	576	(927)
6 cm	G3ZEZ-SM6HYG	July 12, 1983	ducting	610	(981)
3 cm	I0SNY/EA9-I0YLI/IE9	July 8, 1983	ducting	1032	(1660)
12 mm	I3SOY/3,IW3EHQ/3-I4BER/6, I4CHY/6	April 25, 1984	LOS	180	(289)
6 mm	DJ1CR-DL3ER/P	June 11, 1984	LOS	9.3	(15)
10 μm	WA2GFP/2-K2KXS/2	June 10, 1983	LOS	0.2	(0.3)

Note 1. 6 meters has been omitted from this listing because long-path QSOs (those exceeding 12,440 miles or 20,016 km) have been reported during solar cycles 19 and 21.

obtain in this initial phase; I'm not aware of any other attempt to compile and list it all in a single source. I've been filing much of this type of information for over 15 years. Lately, dozens of letters had to be written, and it took the effort of many others to bring this information together, especially on the various propagation modes. In this first attempt, some of the listings may not really represent the best or most recently attained North American record. Because information may, in some cases, be simply unavailable, some propagation modes may not be listed.

I'll be glad to act as a coordinator for all North American and worldwide VHF/UHF/SHF DX claims and will continue to compile and update these records and make them available to ham radio *and other publications.* If you think that you or someone else holds a better DX record than any of those shown in these tables, I'll be glad to consider your claim. For record keeping purposes, I've prepared a comprehensive form to be filled out when claiming a record. Just send me an SASE, appropriately marked, and I'll send you a copy.

propagation breakdown

In last year's propagation column I listed over 20 distinctly different modes of VHF/UHF/SHF propagation. I also made several propagation predictions that came to pass soon after the issue appeared. This experience only reinforced my feeling that the VHF/UHF/SHF frequencies are a great place for experimentation. In light of other information now available, new material should be added to the content of last year's column.

F2 propagation. There's no doubt that solar cycle 21 peaked higher than expected, but it is now approaching minimum. **Figure 1** summarizes sunspot data for the last few cycles. While it may be easy to conclude that the sunspots will bottom out in 1986-1987, they will increase again and probably peak in 1990. However, it is my belief that cycle 22 won't equal the high peak of cycle 21.

I believe this because there seems to be more than enough evidence to link high sunspot activity to the lineup of the planets in certain special arrangements. The late John Nelson of RCA was a strong proponent of this theory and explained it well.[2] The planets line up best about every 175 or so years, with the best alignment occurring in 1984. I doubt that many of us will be around in the year 2159!

Nelson also pointed out that the actual sunspot activity peak occurs when certain major planets are at 90

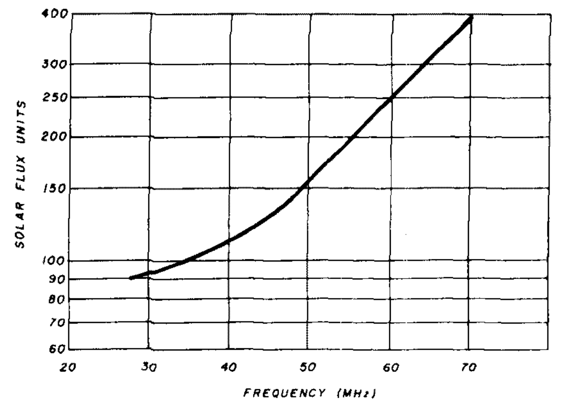


fig. 2. This graph shows the estimated minimum solar flux units necessary for propagation of frequencies from 28-70 MHz per information generated by G8KG.[3]

table 3. Worldwide claimed EME DX records. (See *table 4* for North American EME records.)

band	record holders	date	DX miles	(km)
6 meters	K6MYC-K8MMM	July 24, 1984	2127	(3422)
2 meters	K6MYC/KH6-ZS6ALE	February 18, 1983	12088	(19450)
135 cm	K1WHS-KH6BFZ	November 17, 1983	5058	(8139)
70 cm	F9FT-ZL3AAD	April 18, 1980	11679	(18793)
23 cm	PA0SSB-ZL3AAD	June 13, 1983	11595	(18657)
13 cm	PA0SSB-W6YFK	April 5, 1981	5491	(8836)
9 cm and above	none reported			

degrees to each other with respect to the sun. Furthermore, the minor planets (in particular Mercury and Venus) introduce secondary peaks on the main curve. This probably explains why the F2 activity seemed to peak in 1979, then disappeared and finally came back in 1981 at a slightly diminished level.

Much was learned about F2 on 6 meters during cycle 21. For instance, based on tests between G3SSD and VE1AVX, F. M. Smith, G8KG, has speculated that the 10.7 cm (2800 MHz) solar flux must reach at least 160 for the MUF to reach 50 MHz.[3] Values for other MUFs with equivalent sunspot numbers are shown in **fig. 2**. The 10.7 cm solar flux as measured at Ottawa (the reference station for NOAA) is broadcast at 18 minutes after each hour on WWV and is available any time by calling 303-497-3235.

I have used MINIMUF[4] to predict openings over paths as long as 6000 miles (9654 km) with reasonable accuracy. Solar flux can be determined by using the following approximate equation.

$$\text{solar flux} = 63.7 + 0.73R + 0.0009R^2 \qquad (1)$$

where R is the daily sunspot number.

With this knowledge, improved equipment, and an increase in countries that should have 6-meter privileges by 1990, we should all have something to look forward to during the next solar cycle.

E_s (mid-latitude sporadic-E). Sporadic-E propagation is one of the main propagation modes used by 6-meter operators. In the mid-northern latitudes it usually begins in May and ends in early August. A secondary but weaker peak may come during December and early January.

Ernest Smith and Edwin Davis have been studying E_s propagation for many years.[5] They speculate that E_s propagation is caused when the upper atmosphere, ionized by solar radiation, is subjected to a wind shear. They note that the effects are masked in some parts of the world by precipitation of charged particles at high latitudes and unstable plasmas in the charged particle stream at the magnetic equator. They also point out that during the E_s season, propagation will occur about 1 percent of the time in the southern USA and will drop to 0.3-0.4 percent in the northern USA and southern Canada. Contrast this with a probability of 5-6 percent for the Japanese!

The early 1984 E_s season started off with a bang in early May. Then came a 6-meter opening in mid-May that possibly involved E_s and TE linkup and gave many east coast USA stations their first contacts with Argentina and Chile. Several stations checked 10 meters during this opening, but no signals were heard!

W6JKV made a trip to Nuuk, Greenland in mid-June, 1984. He made about 250 contacts to the USA in widely scattered directions and completed five QSOs with the UK. But by late June openings seemed to come to a screeching halt. One narrow-path W1/W2 to GJ3YHU opening did occur on June 30, but it did not extend into the UK proper (yes, I know there were UK stations who heard GJ3YHU working the USA, but they couldn't hear the USA stations). HB9QQ reported that many 2-meter contacts were made over a wide area of Europe during this same opening.

E_s 6-meter openings returned in late July. As predicted in last year's column, there were some scattered 2-meter openings in late July, but they were gone by early August and few were noted in the December time frame.

There is speculation that a lightning storm whose top reaches an altitude

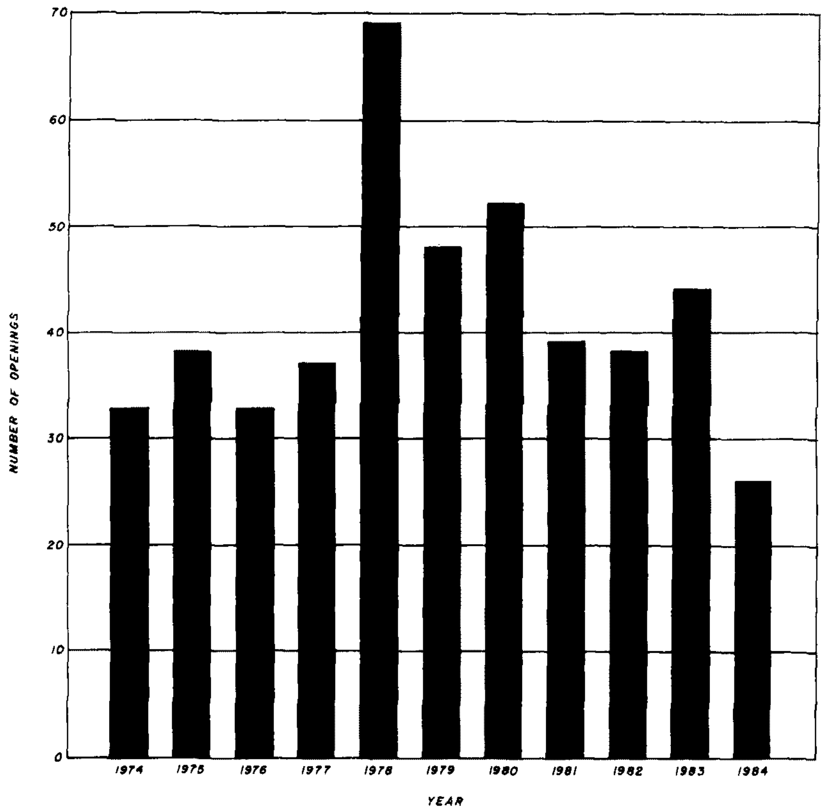


fig. 3. Plot of the observed E_s openings noted in the last 10 years in the UK by Ron Ham.[8]

of 50-60,000 feet (18,288 meters) near the center or at one end of the path may cause a 2-meter opening.[6,7] Such an opening did appear during at least one 2-meter opening in 1984, with a storm (recorded on FAA weather maps) near one end of the path. Jim Stewart, WA4MVI, has seen storm tops as high 72,000 feet (21,946 meters) indicated on the same weather maps!

A mid-December 2-meter opening also occurred between El Paso, Texas, and VE6/VE7. WA4MVI feels that this opening was caused by a special horizontal wind shear force, not as high as the summer ones noted above, but such as typically occur in December in the regions above the Rocky Mountains. FAA weather maps generated at the time of this opening did show an upper-air wind shear at the approximate mid-point of that path!

E_S propagation for the 1984 season as a whole was significantly down from previous years, especially during the winter peak season. Likewise, 6-meter double-hop openings were few in number, especially to the Caribbean. Some Amateurs have speculated that E_S propagation is more intense during low sunspot years, but I can't find any data to substantiate this.

Recently, Ron Ham released his summary of the E_S openings he observed in the UK for 1984 and the 10 years before.[8] I have plotted his observations in **fig. 3**. This data clearly shows the increase in numbers of E_S openings during the high sunspot years.

WA4MVI indicates that the taller thunderstorms, which usually influence the E_S propagation, are more prevalent in years when sunspot activity is high! Therefore, for high E_S activity, I'd say that we may have to wait until the sunspots increase again.

In last year's column I noted that double-hop 2-meter openings have been reported in other parts of the world but not in the USA. I stand corrected; there *have* been some here — of special interest were the ones on 12 July 1982. During this fantastic opening VE1SPI was operational from St. Paul Island, a separate DXCC country in the St. Lawrence River. VE1SPI made about 250 2-meter contacts. Of note is that the operator, VE1ASJ, reported that he could *clearly surmise* double hop: for example, first only W8's and WØ's were heard, then only W3's and W9's, etc.

VE1UT in New Brunswick noted a similar pattern during this opening. Although I may not have located the best DX to occur during that opening (write to me if you can top this), the longest documented 2-meter contact, listed in **table** 4, was clearly a double-hop QSO. With a little bit of luck, we may someday see coast-to-coast 2-meter openings!

Sporadic-E propagation is surely bad news to TV and FM stations. In Europe the lower-frequency TV stations are slowly being replaced by VHF assignments and in the UK all of the low-band TV channels are now silent. This will make it more difficult to observe long DX openings by monitoring European video carriers as we did during the last solar cycles.[1]

However, there are now more 6-meter beacons worldwide with recent additions including one in the UK (GB3SIX) and another in Greenland (callsign unknown). They complement the ZB2BL, FY7THF, KH6EQI and KG6JIH beacons, most of which operate between 50.0-50.1 MHz.

The UK has now licensed over 100 Amateurs to operate on 6 meters outside of the TV hours on the continent. Norway has also licensed at least 25 Amateurs with the same provisions. The lower TV channels in the UK are all gone now and the Norwegian TV assignments in this spectrum are scheduled for shutdown by 1986. Hence, the chances of European DX

on 6 meters are improving, and more countries are looking at Amateur assignments.

Just as this manuscript was going to press, I received a note that Sid Lieberman, WA2FXB, has developed a method to predict possible E_S openings using the K indices from WWV.[9] Perhaps he'll be able to shed some light on the prediction of E_S openings. E_S propagation is still widely studied and we may someday be able to predict it with good accuracy.

T.E. scatter. Trans-equatorial scatter has not been too common lately except on 6 meters in the equinoctial time periods. Hopefully more work can be done when cycle 22 begins and the necessary solar activity reappears. Likewise, equatorial FAI (field aligned irregularities), ionospheric scatter, and midlatitude FAI propagation are lower in these low sunspot years but they will return! **Table 4** shows that within the USA there has been some real 2-meter DX via midlatitude FAI.

Aurora. This mode of propagation is also heavily dependent on solar activity, and in particular, solar flares. I've been told that the incidence of auroral propagation increases with solar activity, but that the greatest number of auroras appears when the sunspots are declining.

This is why I keep careful notes in my logs on known auroras. Sure enough, there's a definite trend. **Figure 4** shows a plot of the number of auroral openings I've observed over the last eight to ten years. Note the increase in auroral openings as the sunspots decreased in 1982.

G2FKZ has plotted auroras since 1932.[10] He notes that the highest incidence of aurora on a month-by-month basis occurs in April, September, and October, in that order. December, January, February, and November are significantly lower in activity, with only about one-fourth the occurrences of April, September, and October. (See **fig. 5**).

I've been told that Canadian Research Labs (CRL) has done a lot of

table 4. North American claimed VHF/UHF/SHF DX records, listed alphabetically by the most common modes of propagation. The (tropo) ducting records are for paths that are mostly over water. (See text for how these records were determined and how you can challenge or add to those records shown.)

band	record holders	date	mode	DX miles	(km)
2 meters	K2RTH-W5WAX	March 8, 1970	aurora	1221	(1964)
	VE1UT-VK5MC	April 7, 1984	EME	10985	(17,676)
	K0UDZ-VE1UT	July 12, 1982	E_s	1832	(2947)
	W5HUQ/4-W5UN	July 25, 1983	FAI	1229	(1977)
	K1ABR-W5ORH	August 12, 1968	MS	1469	(2364)
	KP4EOR-LU5DJZ	February 12, 1978	TE	3933	(6328)
	K1RJH-K5WXZ	October 8, 1968	tropo	1465	(2358)
	KH6GRU-WA6JRA	July 29, 1973	ducting	2591	(4169)
135 cm	W1FC/1-W0VB	June 13, 1982	aurora	1039	(1672)
	K1WHS-KH6BFZ	November 17, 1983	EME	5058	(8139)
	WB4NMA-W5FF	August 12, 1983	MS	1273	(2048)
	KP4EOR-LU7DJZ	March 9, 1983	TE	3670	(5906)
	VE3EMS-WB5LUA	September 28, 1982	tropo	1181	(1901)
	KH6UK-W6NLZ	June 22, 1959	ducting	2540	(4087)
70 cm	K1PXE-W0RAP	July 13, 1982	aurora	957	(1540)
	K2UYH-VK6ZT	January 29, 1983	EME	11567	(17,612)
	W2AZL-W0LER	August 12, 1972	MS	1020	(1641)
	WA2LTM-WB5LUA	September 10, 1979	tropo	1310	(2108)
	KD6R-KH6IAA/P	July 28, 1980	ducting	2550	(4103)
23 cm	K2UYH-VK5MC	December 6, 1981	EME	10562	(16,995)
	WA4OFS-W5VY	March 25, 1985	tropo	1046	(1683)
	KH6HME-N6CA	June 24, 1984	ducting	2472	(3977)
13 cm	PA0SSB-W6YFK	April 5, 1981	EME	5491	(8836)
	W4HHK-W8YIO	July 28, 1983	tropo	583	(938)
9 cm	K6HIJ/6-W6IFE/6	June 18, 1970	LOS	214	(344)
6 cm	K5PJR-K5FUD	September 20, 1977	tropo	267	(430)
3 cm	W7JIP/7-W7LHL/7	July 31,1960	LOS	265	(426)
	WA4GHK/4-WD4NGG	August 7, 1984	ducting	297	(478)
12 mm	W2SZ/1-W2JVF/2	September 8, 1984	LOS	53	(86)
6 mm	W2SZ/1-WA2AAU/1	June 13, 1982	LOS	0.3	(0.5)
10 μm and up	WA2GFP/2-K2KXS/2	June 10, 1983	LOS	0.2	(0.3)

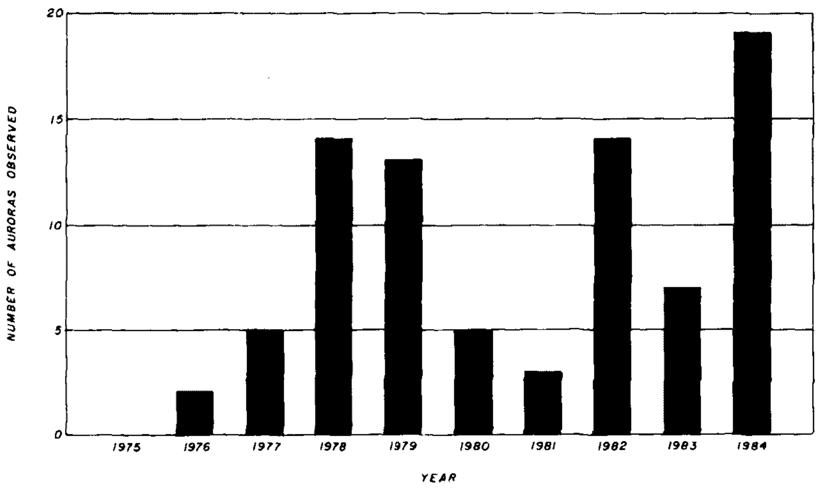


fig. 4. Plot of the number of observed auroral openings at W1JR for the last 10 years.

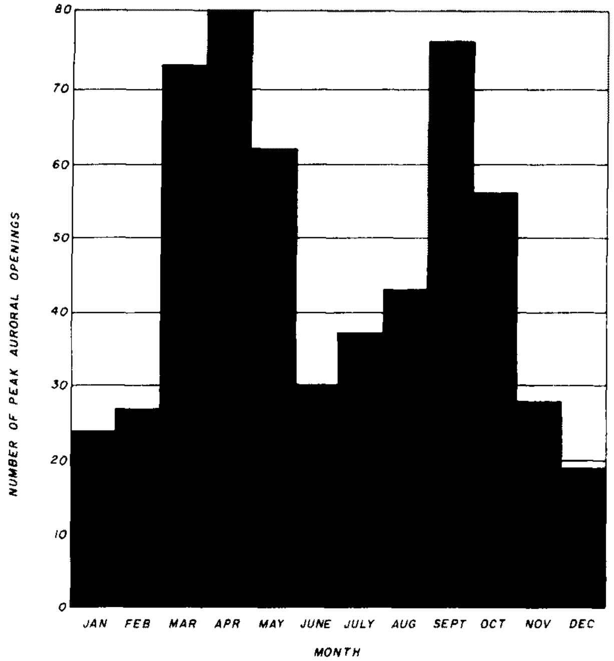


fig. 5. Peak auroral openings by months since 1932 from data by G2FKZ.[10]

aurora research, but I haven't yet received any of their papers. CRL has noted that the auroral oval stays further north during low sunspot years and extends southward during years of high sunspot activity.[1] This would explain why some of the southern portions of the United States, which rarely see auroral propagation, were treated to some good openings during the peak of cycle 21. Radio aurora was also the subject of a recent *QST* article.[11]

In summary, auroras may decrease in frequency during the next few years but they will not disappear entirely. Look for a WWV *K* index of 5 or greater for an early indication of aurora.[1] If you plot the WWV *K* indices, you'll find that auroras often return 28 or so days later. NOAA's weekly report is also valuable in summarizing past data and predicting future sunspot, *A* and *K* indicies.[12]

Meteor scatter. This is "the fun mode" that heretofore wasn't thought to have been affected by solar activity.

Some Swedish scientists, however, now report that from their observations, meteor scatter propagation seems to improve during the period of low solar activity.[13]

For instance, meteor counts made by radars in Sweden were higher by a factor of 2 during the sunspot minimum in 1963 than during the maximum in 1956-57. They also found that while the beginning heights of meteor trails varied little from year to year, the terminal heights were 6.84 miles (11 km) higher at sunspot minimum. Hence the meteor showers during the next few years should improve. Couple this with the possibility of a link between Halley's comet (arriving in late 1985) and some of the major meteor showers (such as the Eta Aquarids and the Orionids) and we could see some superior meteor shower performance to offset other sunspot-related propagation modes.

The amount of data available strongly supports the technique for predicting meteor shower peaks described in my June, 1984, column.[14] In fact, this method predicted the 1985 Quadrantids shower peak to within hours of its beginning. But there seems to be some disagreement on the accuracy of using the ecliptic longitudes as shown in **table 2** in that article. Due to slight shifts in the earth's orbit, errors of a few hours will gradually creep into the ecliptic longitude at 0000 UTC as the years go by.

However, the use of the computer program provided in **fig. 1** of the July, 1984, article is thought to be a more accurate method for predicting the peaks. This all may be somewhat academic because most showers last 24 to 48 hours during peak (greater than 25 percent of maximum), so **table 2** should be OK for the next few years.

In the same column, I made a negative comment about packet radio meteor scatter contacts. Since that article appeared, the first documented contacts have taken place on 6 and possibly 2 meters. My remarks were made in a tongue-in-cheek fashion as a joke with one of my packet radio friends, Jeff Moore, KQ1E. Those who know me know that I'm never going to stand in the way of progress . . . but there's something about actually *hearing* and completing a meteor scatter QSO. For some, this means seeing a message appear on a CRT! Go to it. I won't stand in your way!

Meteor scatter communications is one of the most important propagation modes for the VHF/UHFer. More work has to be done, especially on 135 and 70 cm. There are many unused showers available for exploration. The use of the VHF/UHF calling frequencies is a good step toward increasing random contacts. Home computer programs for predicting meteor shower peaks as well as optimum direction and time of day are very powerful tools.

EME. This mode of propagation has really taken off. In 1984, there were reported contacts on 6 and 2 meters as well as on 135, 70, 23, and 13 cm! Techniques and equipment are steadily improving. Low-cost GaAs FETs are now available that will deliver the ultimate in low noise figures required for EME.[15]

Most of the EME action is on 2 meters and 70 cm. Some of the larger 2-meter stations have huge arrays,[16] which allows smaller stations, with a

single Yagis and moderate power (500 watts output), to routinely make contacts with the larger stations using a rising or setting moon.

Although the 135-cm band has been very much ignored by the EMEers, it's a terrific band for EME. Antenna systems are not as critical and are only 50-75 percent of the size required for 2-meter EME. Because the sky noise is lower, the signals are stronger. All the necessary 135-cm EME components can now be built or purchased.[17] *Warning: this is a band we could lose if we don't start using it properly.*[18]

EME on 70, 23, and 13 cm is maturing rapidly. Over 40 DXCC countries are available on 70 cm on all continents. Twenty DXCC countries from five continents are now active on 23 cm and almost ten countries are on 13 cm. On 23 and 13 cm, the parabolic dish is king. If low-noise GaAs FET preamplifiers are used, echoes can be obtained with as little as 100 watts and a 13-foot (4 meter) parabolic dish.

The stout-hearted and others in search of a real challenge should give EME a try. Still quite an adventure, it's an excellent proving ground for new equipment. I think we'd be at least five years behind in antenna technology today if it weren't for the need for high-gain efficient Yagi antennas required for EME. Now these same antennas are being used to improve performance on other, more conventional, propagation modes.

Weather-related propagation. Watching the weather and trying to predict it was a hobby of mine until I discovered Amateur Radio. This background has served me especially well since I began working the VHF/UHF frequencies.

Over the years I've noticed a strange weather phenomenon, especially in climates that experience temperature changes of greater than 75 degrees F (42 degrees C): *there seems to be more than four distinct seasons.*

According to my "Five-Season Weather Calendar," the weather appears to change on five key pivot calendar dates, 73 days apart. I find that March 1 is a good reference point for determining the first pivot date. The other dates are May 13, July 25, October 6, and December 18.

What I notice is that the weather in any particular area of the country around each pivot date (± 1 week) gives an indication of the weather to be expected during the next 73-day period. For instance, if the weather is generally cold at the end of February through the first week of March, the weather to follow will probably be cool until the next pivot day (May 13). Likewise, in the northern hemisphere, moderate temperatures near the weeks surrounding December 18 portend a milder winter season.

Furthermore, this five-season weather concept seems to be in step with radio propagation. For instance, the E_S season seems to begin around May 13 and is usually almost over by July 25. Could it be that the jet stream moves in approximately 73-day increments? Hurricanes in the United States seem to form in late July and are usually over by early October. Is this concept only a figment of my imagination? Or is it really so?

Tropospheric propagation. There's no doubt that tropospheric propagation, tropospheric ducting, and super-refraction are directly related to the weather patterns. As mentioned in reference 1, long-haul tropo DX seems to come after a slow-moving high-pressure area (greater than 30.27 inches of 1025 millibars on a barometer) is followed by a moist low-pressure system. Recently, VE3CIE wrote an interesting article on tropo propagation as related to meteorology.[19]

The past year has seen some good long tropospheric openings both in North America and Europe. First there was the ducting from California to Hawaii when KH6HME and N6CA finally made the grade on 23 cm. At the time a tropical depression was noted on the southern side of the path off Baja, California. As in the past openings, the signals disappeared a few miles inland at the California end of the path.

Next came the openings between the Canary Islands (EA8) and the United Kingdom in July. This opening was also a function of ducting, since only stations near the coast of southern England and Wales were able to hear the EA8 stations. Signals were strong and relatively small stations (10 watts) were able to make contacts.

Finally came the fantastic opening that extended from New England and New York to Florida during the ARRL September VHF QSO party. This was a classic opening that was clearly a mix of normal tropo and ducting. The barometer was very high — and right off the east coast was the large hurricane Diana! Who says there's no link between hurricanes and good DX?

The salient feature of this opening was the evidence of an elevated duct. The stations located at least 500 feet (150 meters) above sea level were particulary favored. W2SZ/1 and W1XX/2 were over 3000 feet (900 meters) above sea level and they reported that making long-haul contacts was like shooting fish in a barrel. W2SZ/1 operators noticed a distinct haze layer above and below their mountaintop location in the early morning.

WA4MVI, a private pilot, decided to observe the opening from the air. He loaded his 70-cm gear, including a five-element Yagi, into a small plane. He then flew at various elevations over the western tip of South Carolina and into North Carolina to see if there was a duct and, if so, where it was located. At 750 miles (1200 km) from W2SZ/1, he was able to continuously monitor their 70-cm signal strength from approximately ground level to about 14,000 feet (4,267 meters). He found a duct between about 4000 feet (1200 meters) and 10,000 feet (3050 meters). Signals in the duct were typically 20 dB over S9. However, signals abruptly weakened above and below the duct, dropping to almost inaudible at ground level and at 14,000 feet. His temperature-versus-elevation observations showed a more or less constant value within the duct instead of the normal decrease with increasing altitude. He also noted that the wind speed in the

duct was very high — typically 40 knots — and that the wind came from the east. However, above and below the duct there was only a moderate 10-knot wind — coming from the *west*!

In mid-December, when such phenomena are rare, another east coast tropospheric duct period was in evidence. Also present, just off shore, was an unwelcome guest — hurricane Lily! (The weather reports said that this was one of the few hurricanes ever seen in that part of the Atlantic Ocean during the month of December.)

And how about the terrific openings during the ARRL VHF contest in September, 1979, when hurricanes David and Frederic were both off the Southeastern coast of the United States? Again, the link between hurricanes and good long-haul VHF/UHF DX seems difficult to dispute!

Openings such as this prompt some hams to remark "Oh, the bands are always open. It's just that we don't have all that mountaintop activity outside of the contests." *This just isn't so*. Most of the VHF/UHF contest dates are planned to coincide with periods known to have good radio propagation. Mountain-toppers are always active, especially during contests, but they usually experience only the normal extended range expected for an elevated QTH. To have an extended opening, you have to have the right ingredients: proper weather *and* location.

In my July, 1984, column I stated that ducts do not extend far inland. This statement applies mainly to the oceans, especially when the land near the coast is hilly or mountainous. The Gulf of Mexico does not fit this description but instead acts more like a large inland body of water surrounded by relatively flat land. As a result, ducting *can* move inland.

This is most noticeable on the path between Texas and Florida, especially between the months of February and May. Just as this column was being completed, I learned of a particularly long DX contact (approximately 1046 miles or 1683 km) on 23 cm, between WA4OFS (St. Cloud, Florida) and W5VY (San Antonio) — see **table 4**. Both stations are very far inland.

The next few years should prove very interesting as the jet stream moves slightly because of lower sunspot counts. Will tropo propagation disappear? I doubt it very much — but it may favor different areas of the country than it does during the high sunspot years.

Lightning scatter. Over the past year I've had many reports from individuals who've used various scatter mechanisms. W7BYF informed me that he made what was undoubtedly a lightning scatter contact when he was W8NAF in Dayton, Ohio. In July of 1958 he worked W8KAY in Akron, Ohio, on 2 meters, with both stations pointing their antennas at a lightning storm center in Ft. Wayne, Indiana. Several other stations also worked W8KAY by the same technique.

To work lightning scatter, set up a schedule for 15-30 minute periods with a distant station. Each station transmits for 1 minute — one on the odd minute and the other on the even minutes. Listen carefully when working extended paths. More often than not, the signals will seem to appear almost out of nowhere. When signals appear, switch over to break-in procedures and enjoy a quick snappy QSO. The better the location and the shorter the distance between stations, the longer the propagation will last.

Aircraft scatter. VHF/UHFers often seem surprised when you mention that they may be using aircraft scatter propagation. However, scatter is often present on 200-400 mile (325-650 km) contacts even though it may not be obvious on the lower VHF bands. On 70 cm and above, aircraft is most often responsible for extended daily contacts. Using aircraft scatter requires some patience, since aircraft may be in the proper location for only a few minutes.

Barium clouds. The jury is still out on whether VHF/UHFers can use these man-made ionization clouds, which apparently last for 15 to 30 minutes, for comunications. On Christmas morning, 1984, there was supposed to be a Barium cloud, dubbed the Christmas Comet, released over Peru at 70,000 miles (21,336 km) altitude. Unfortunately the test had to be rescheduled, so many persons missed their chance to try, and those who did were apparently unsuccessful. These opportunities, which occur infrequently and often unexpectedly, deserve more attention.

finding direction

When working DX on the UHF bands, your antenna can have very narrow *beamwidths*. This *requires* good rotator accuracy as well as a knowledge of the correct beam headings. One way to calibrate your rotator in the northern hemisphere is to aim your antenna at Polaris (the North Star).

You can also use the sun. By consulting your local daily newspaper for times of sunrise and sunset, you can estimate the time the sun passes directly south. Simply determine the midpoint between sunrise and sunset. For instance, if the sun rises at 5:30 AM and sets at 7 PM local, the time of southerly transit or time when the sun is directly south will be 12:15 PM.

A compass can also be used. But beware — magnetic north may be different than true north. Since many VHFers now use home computers and have direction bearing programs, it is easy to find the true bearing for magnetic north. For USA stations, just compute the direction of the magnetic north pole using the approximate coordinates of 74 degrees north latitude and 101 degrees west longitude. For southern hemisphere stations, the southern magnetic pole is at approximately 68 degrees south latitude and 144 degrees east longitude. This will give you the true direction indicated on a hand compass.

summary

Again, I've run out of time and space, but I hope the information presented will be useful. If you've never seen my July, 1984, column, I suggest

you obtain a copy of that issue, since this month's column is based on it to a large degree.*

Radio propagation is a fascinating science, and one that can be advanced by the Radio Amateur. New equipment, increased activity, more propagation beacons, and closer attention to possible openings has greatly helped.

acknowledgements

I'd particularly like to thank all those who helped me with encouragement and material on their propagation observations. Special thanks go to Jim Stewart, WA4MVI, for all his help. Thanks also to the many others who helped assemble the DX records, especially: KP4OER, PAØSSB, SM5AGM, VE1UT, K1WHS, WA2SPL, K2UYH, W5FF, W5HUQ, WB5LUA, K6MYC, K8MMM, and ZL3AAD.

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*Back issues are available for $3.00 each from *ham radio*, Greenville, N. H. 03048

REMINDER: Send an SASE for a DX record claim form if you want to claim any improved records. Address your request to me, Joe Reisert, W1JR, 17 Mansfield Drive, Chelmsford, Massachusetts 01824.

upcoming VHF/UHF events

July 20:	*Look for 2-meter E_s openings ± 2 weeks*
July 20-21:	*CQ Magazine VHF WPX Contest*
July 25:	*EME Perigee*
July 27-29:	*Central States VHF Conference, Tulsa, Oklahoma (WØRRY/5)*
July 28:	*Predicted peak of the Delta Acquarids meteor shower (0300 UTC)*
August 3-4:	*ARRL UHF Contest*
August 12:	*Predicted peak of Perseids meteor shower (0130 UTC)*
August 20:	*EME Perigee*

ham radio

DX FORECASTER

Garth Stonehocker, K0RYW

summertime DX

On the higher frequency bands, 6 through 30 meters, summertime DX is usually very good. Between sporadic-E short-skip openings around noontime and the bands opening earlier and staying open longer, there's plenty of DX fun to keep us happy all day and well into the evening. On 80 and 160, however, received signal levels will be lower because of increased ionospheric absorption, and static will be higher.

This QRN, propagated from the equatorial land regions, increases the overall average noise level of the lower HF bands, peaking at about 10 MHz, dropping somewhat, and then rising sharply just below 4 MHz.

At any given moment, an estimated 3600 thunderstorms are in progress around the world. They can be classified as air-mass, frontal, or orographic, depending on how they are formed. Some are combinations of the three types — that is, an air-mass or frontal storm crossing a mountain range may rise with the change in elevation to become an orographic storm.

Some regions of the country have a greater number and variety of thunderstorms than others. This is measured by the number of thunderstorm days per year. (A thunderstorm day is a day in which at least one storm occurs.) Areas with 100 thunderstorm days or more are found in Florida and in the Rocky Mountains; the southern parts of Louisiana, Alabama, and Georgia see 80 days of thunderstorm activity. A band stretching across Nebraska to Ohio and then bending southward to North Carolina, and another reaching across New Mexico to Northern Texas, see 60 days, and the midwestern states between these bands experience 50 thunderstorm days. The main source of summertime QRN is the air-mass thunderstorm, which builds up from the sun's heating the ground and the air above it.

Most air-mass storms form in afternoons when the humidity is above about 50 percent, and last into the night before cooling off enough to dissipate. Unlike spring or fall frontal passage thunderstorms, which simply pass by, air-mass thunderstorms linger for several days until rain releases their moisture or they slowly move on. During the evening DXing hours air-mass thunderstorm QRN may limit the usefulness of low-band signals to local ragchewing and rule out weak-signal DX.

So how do you get some DXing in on these bands? Most operators switch operating hours, giving up evenings in favor of the pre-dawn hours of early morning. By this time, the thunderstorms have dissipated to the east, locally, and are dissipating on paths to the west. This is a cool, comfortable time of the day to be up and around.

last-minute forecast

The best opportunity for good DX conditions on the higher frequency bands, 10 through 20 meters, will occur during the last week and a half of the month, when the solar flux may be a little higher. (Listen to WWV at 18 minutes after the hour for the solar flux. Any value above 80 is high.) Six meters can have openings anytime during the month. The middle of the month will favor working short-skip both day and night on the lower frequency bands. Disturbances are more likely around the 1st, 9th, 18th, and 28th days of July.

A full moon will occur on the 2nd and 31st; perigee (closest approach of the moon) is on the 25th. The Aquarids meteor shower begins on July 18, peaks on the 28th, and lasts until August 7. (All dates are approximate, but close.) The radio-echo rate at maximum is about 34 per hour.

band-by-band summary

Six-meter paths will open for a half hour to a couple of hours on some days around local noon. Sproadic-E propagation will make this short-skip path possible out to nearly 1200 miles (2000 km) per hop.

Ten, fifteen, twenty, and thirty meters will support DX propagation from most areas of the world during daylight and into the evening with long-skip out to 2000 miles (3500 km) per hop. Sporadic-E short-skip will also be available on many days for several hours near local noon. The direction of propagation will follow the sun across the sky: morning to the east, south at midday, and west in the evening. Long daylight provides many hours of good DXing. Solar flux is so low this year that daytime absorption allows higher signal strengths than usual on these bands during this month.

Thirty, forty, eighty, and one-sixty meters are the nighttime DXer's bands. On many nights 30 and 40 meters will be the only usable bands because of thunderstorm QRN. Try the pre-dawn hours for the best DX. The direction of propagation follows the darkness path across the sky: evening to the east, south around midnight, and toward the west in the pre-dawn hours. Distances will decrease to 1000 miles (1600 km) for skip on these bands. Sproradic-E openings will be most frequently observed around sunrise and sunset. These may be the only signals getting through the noise in the evening. Again, because of low solar flux, daytime DX — particularly in the mornings — may be good this month.

JULY

WESTERN USA

GMT	PDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	5:00	20	20	15	15	20	10	15	20
0100	6:00	20	20	20	15	20	10	15	20
0200	7:00	20	20	20	15	20	10	15	20
0300	8:00	20	20	20	15	20	15	15	20
0400	9:00	20	20	20	15	20	15	15	20
0500	10:00	20	20	20	15	30	15	15	20
0600	11:00	20	30	20	20	30	20	15	20
0700	12:00	20	30	30*	20	30	20	15	20
0800	1:00	20*	30	30	20	30	20	15	20
0900	2:00	20*	30	20	20	30	20	20	20
1000	3:00	20	30	20	20	30	20	20	20
1100	4:00	20	30	20	20	30	20	20	20
1200	5:00	20	20	20	20	40	20	20	30
1300	6:00	20	20	15	30*	40	20	20	30
1400	7:00	20	20	15	30*	30	20	20	30
1500	8:00	20	20	15	20	30	20	30	30
1600	9:00	20	20	15	20	20	20	30	30
1700	10:00	20	20	15	20	20	15	20	30
1800	11:00	20	20	15	15	20	15	20	20
1900	12:00	20	20	15*	15	20	15	20	20
2000	1:00	20	20	15*	15	20	15	15	20
2100	2:00	20	20	15	15	20	15	15	20
2200	3:00	20	20	15	15	20	15	15	20
2300	4:00	20	20	15	15	20	15	15	20

MID USA

GMT	MDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CDT
0000	6:00	20	20	15	15	20	10	15	20	7:00
0100	7:00	20	20	20	15	20	10	15	20	8:00
0200	8:00	20	20	20	15	20	10	15	20	9:00
0300	9:00	20	20	20	15	20	15	15	20	10:00
0400	10:00	20	20	20	15	30	15	15	20	11:00
0500	11:00	20	20	20	15	30	15	15	20	12:00
0600	12:00	20	30	20	20	30	20	15	20	1:00
0700	1:00	20	30	30*	20	30	20	15	20	2:00
0800	2:00	20	30	30	20	30	20	15	20	3:00
0900	3:00	20	30	20	20	30	20	20	20	4:00
1000	4:00	20	30	20	20	40	20	20	20	5:00
1100	5:00	20	30	20	20	40	20	20	20	6:00
1200	6:00	20	20	20	20	40	30*	20	30	7:00
1300	7:00	20	20	15	30*	40	30*	20	30	8:00
1400	8:00	20	20	15	30*	30	20	20	30	9:00
1500	9:00	20	20	15	20	30	20	30	30	10:00
1600	10:00	20	20	15	20	30	20	30	30	11:00
1700	11:00	20	20	15	20	20	15	20	20	12:00
1800	12:00	20	20	15	15	20	15	20	20	1:00
1900	1:00	20*	20	15*	15	20	15	20	20	2:00
2000	2:00	20*	20	15*	15	20	15	15	20	3:00
2100	3:00	20*	20	15	15	20	15	15	20	4:00
2200	4:00	20	20	15	15	20	15	15	20	5:00
2300	5:00	20	20	15	15	20	15	15	20	6:00

EASTERN USA

GMT	EDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	8:00	20	20	15	15	20	10	15	20
0100	9:00	20	20	20	15	20	10	15	20
0200	10:00	20	20	20	15	20	10	15	20
0300	11:00	20	20	20	15	30	15	15	20
0400	12:00	20	20	20	15	30	15	15	20
0500	1:00	20	20	20	15	30	15	15	20
0600	2:00	20	30	20	20	30	20	15	20
0700	3:00	20	30	30*	20	30	20	15	20
0800	4:00	20	30	30	20	30	20	15	20
0900	5:00	20	30	20	20	30	20	20	20
1000	6:00	20	30	20	20	40	20	20	20
1100	7:00	20	30	20	20	40	20	20	20
1200	8:00	20	30	15	20	40	30*	20	20
1300	9:00	20	20	15	30*	40	30*	20	20
1400	10:00	20	20	15	30*	30	20	20	20
1500	11:00	20	20	15	20	30	20	30	20
1600	12:00	20	20	15	20	30	20	30	20
1700	1:00	20	20	15	20	20	15	20	20
1800	2:00	20	20	15	15	20	15	20	20
1900	3:00	20	20	15*	15	20	15	20	20
2000	4:00	20*	20	15*	15	20	15	15	20
2100	5:00	20	20	15	15	20	15	15	20
2200	6:00	20	20	15	15	20	15	15	20
2300	7:00	20	20	15	15	20	15	15	20

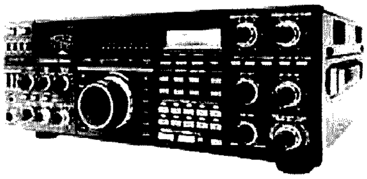
The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

NEW products

Kenwood TS-940S HF transceiver

Kenwood has announced the release of its new HF transceiver, the TS-940S. Designed with the serious HF operator in mind, the TS-940S features all the latest state-of-the-art features. High performance interference rejection circuits and a high dynamic range receiver, combined



with a superior transmitter and excellent audio, provide power for working those really tough ones. Casual DXers as well as contest operators give Kenwood high marks for QRM fighting features such as SSB slope tuning, variable CW bandwidth, notch, AF tune and CW pitch controls.

The TS-940S covers all HF Amateur bands — including the new WARC bands — and is a general coverage receiver for 150 kHz to 30 MHz. Fully self-contained, the unit will operate on FM, SSB, CW, AM, and FSK RTTY; there are no extra modules to buy. The TS-940S has two optically encoded VFOs and features 40 non-volatile memory channels in four groups of ten. You can QSY instantly by using the keyboard on the front panel; just enter the frequency you want. The TS-940S has an LCD sub-display that shows CW, VBT, SSB slope tuning as well as frequency, time and status of the optional AT-940 160-10 meter automatic antenna tuner.

Kenwood offers a full line of optional accessories such as the AT-940 160-10 meter automatic antenna tuner, SP-940S external speaker with audio filtering, a full line of both first and second IF crystal filters so that performance can be tailored to individual tastes, and the VS-1 voice synthesizer, to name just a few.

For more information, contact Trio-Kenwood Communications, 1111 West Walnut Street, Compton, California 90220.

A5 switches focus, changes name, and offers free samples to *ham radio* readers

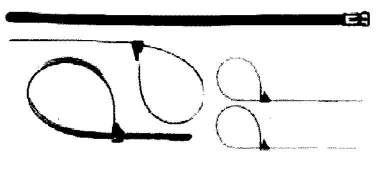
After 18 years of providing in-depth reporting of news and technical developments in Fast Scan TV, *A5* is expanding its range to include coverage of other modes of specialized communications in Amateur Radio. While FSTV will remain the primary object of interest, NBTV, MSTV, SATV, TVRO, SSTV, FAX, RTTY, AMTOR, Packet, OSCAR, EME, and other modes will be addressed in an extended format.

With the change in editorial focus comes a new name: *SPEC-COM.*® *SPEC-COM* will be published ten times a year. One-year subscriptions are available for $20; special six-month trial subscriptions are available to new readers at $10.

For a free sample copy and further details, contact *SPEC-COM — Specialized Communications Journal*, P.O. Box H, Lowden, Iowa 52255. (Allow two to three weeks for delivery.)

cable ties

Available in packages of 20 ($7.00), FLEXLOC reusable cable ties are ideal for bunching cables attached to Amateur Radio equipment, computers and peripherals, home entertainment systems, A/V devices, and other electronic and electrical hardware. Their quick-release ("pinch of the fingers") lock enables repeated re-use. Made from tough, flexible nylon (Type 66), FLEXLOC has a tensile strength of 50 pounds. Each 10-inch fastening strap is self-locking with a ribbed backing that permits diameter adjustments up to 2-3/4 inches and secures the bundle until release.



✓ 167

✓ 166

NEW products

For further information, contact Visual Departures, Ltd., 1641 Third Avenue, Suite 202, New York, New York 10128.

Circle #302 on Reader Service Card.

full half-wave UHF antenna for portables

Larsen Antenna's new KD 14-450-HW antenna, a full half-wave UHF antenna for portables, interfaces with any BNC output portable. Said to offer performance equivalent to a full quarter wave on a perfect ground plane, the KD 14-450-HW is appropriate for any UHF application requiring maximum performance. Because of its inherent resonance, it may be easily remoted via a coaxial line from the portable. The overall whip length is 12 inches; the impedance transformer at the base of the whip is 3.25 inches long.

For more information, contact Larsen Electronics, P.O. Box 1799, Vancouver, Washington 98668.

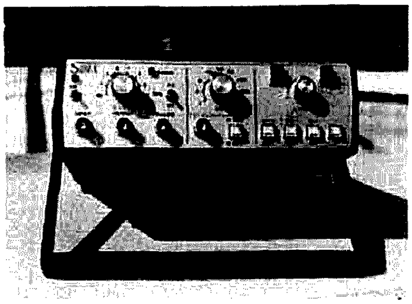
Circle #303 on Reader Service Card.

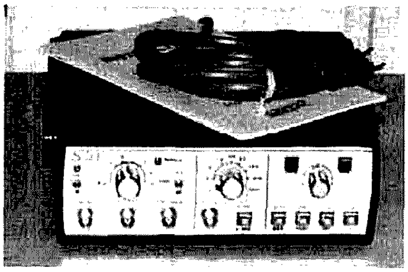
scope memory

Sibex, Inc. has introduced a new device called the Scope Memory. This instrument attaches to your oscilloscope and converts it into a high performance digital storage scope. The Scope Memory can be an economical answer for the service technician, hobbyist, or manufacturing firm that needs a storage scope but cannot justify its expense.

The Scope Memory stores low frequency signals, transients, and one-shot pulses in a single sweep. It stores both analog and digital signals and features 18-selectable sampling times with a 1.4 MHz maximum rate, an 8-range input selector with over-range indication and a wide variety of triggering modes.

The pre- and post-trigger modes allow viewing the waveform that occurs both before and after an event. This feature makes the device a





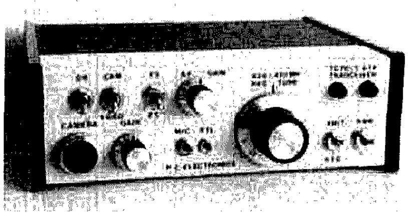
useful tool for trouble shooting. The price is $515.00 plus shipping.

For additional information, contact Sibex, Inc., 2340 State Road 580, Suite 241, Clearwater, Florida 33515.

Circle #304 on Reader Service Card.

ATV transceivers

P.C. Electronics has introduced a new compact 1-watt 70-cm ATV transceiver aimed at introducing hams to the video mode.



The TC70-1 ATV transceiver accepts standard composite video input from any source. Video and audio input RCA jacks on the rear panel are provided for connection to black and white or color cameras, computers, VCRs, and TVROs. A front panel switch selects video and audio input from these jacks or from the 10-pin connector provided for direct connection to many of the popular color cameras made for portable VCRs.

Audio input is selected from the color camera microphone or line level from the rear panel jack. In addition there is a microphone input that accepts any low impedance dynamic microphone. Next to the microphone input is a mini jack for PTL. A "push-to-look" feature resembles push-to-talk on audio-only transceivers, can be used for microphone or remote transmit receive switching or the front panel toggle switch may be used.

Full-color live action video and sound are transmitted with over 1 watt PEP on one or two selected crystal controlled frequencies in the range of 425 to 440 MHz in the 70-cm Amateur band. The line-of-sight snow-free radius with TC70-1s and KLM 440-27 antennas at each end is 15 miles. The unit was made small and compact (7 × 7 × 2.5 inches) for portable use, but either a 20-watt or 50-watt video compensated RF linear amplifier for greater distance, base or mobile, is available.

NEW products

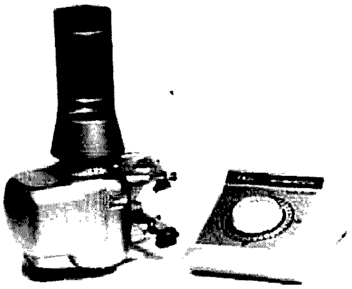
With the TC70-1, the only other items necessary to get on ATV are a good 70-cm antenna, low loss coax, a TV set, and any device with a standard 1 volt P-P composite video output commonly found on black and white CCTV cameras, home video color cameras and VCRs, computers, and RTTY/video converters. A Technician class or higher Amateur radio license is required for purchase of this equipment from P.C. Electronics and subsequent operation.

Priced under $300, the TC70-1 makes getting on this exciting visual mode affordable. For further information, contact P.C. Electronics, 2522 Paxson Lane, Arcadia, California 91006.

Circle #305 on Reader Service Card.

antenna rotator

The AR-200XL antenna rotator operates from 115 VAC and provides 220 pounds/inch torque to turn an antenna array or surveillance camera. Full 360-degree rotation is achieved in 60 seconds. Motor voltages are held below 18 VAC for safety and only three conductors are required between the control unit and rotator. The control unit incorporates a demand heading control and a present heading indicator presented concentrically on a compass rose. Designed for medium-duty applications, the rotator will support a vertical load of up to 100 pounds with a wind loading of 5 square feet.



For further information, contact CMC Communications, Inc., 5479 Jetport Industrial Boulevard, Tampa, Florida 33614.

Circle #306 on Reader Service Card.

new receivers

Yaesu Electronics has announced two additions to its line of high-performance receiver products.

nateur Radio Dealer

Ohio

AMATEUR ELECTRONIC SUPPLY
28940 EUCLID AVE.
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216-585-7388
Ohio Wats: 1 (800) 362-0290
Outside Ohio: 1 (800) 321-3594
Hours M-F 9-5:30, Sat. 9-3

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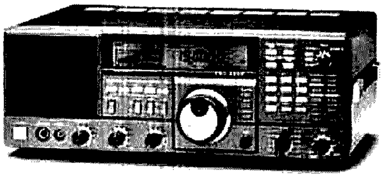
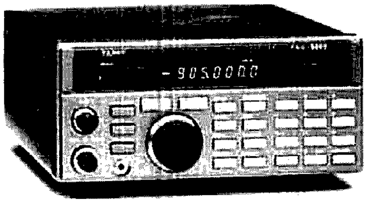
MILWAUKEE, WI 53216
414-442-4200
Wisc. Wats: 1 (800) 242-5195
Outside Wisc: 1 (800) 558-0411
M-F 9-5:30
Sat 9-3

NEW products

The FRG-8800 is a deluxe HF receiver covering 150 kHz through 29.99 MHz. Direct frequency entry is provided via the front panel keyboard, which also controls scanning functions and storage/recall of the memory channels. The innovative green LCD information display provides frequency, mode, and signal strength information. Selectable AGC, all-mode squelch, two 24-hour clocks, and recording capability (including on/off timer switching) make for maximum operating flexibility. The FRG-8800 is designed for easy interface to a personal computer for expanded operating control, and the FRV-8800 VHF Converter option expands coverage to include 118-174 MHz, with front panel entry and display.





The FRG-9600 is a high-tech VHF/UHF multimode receiver covering 60 through 905 MHz continuously, with direct keyboard frequency entry and 100 memory channels. Capable of AM, FM, and SSB reception (SSB through 460 MHz), the FRG-9600 comes equipped with both wide and narrow bandwidths on AM and FM, plus a total of seven different tuning steps from 100 Hz through 100 kHz, depending on the mode and typical operating requirements for the service being received. A variety of scanning features are included, with extra performance offered via the audio scan stop function which will prevent stopping on inactive "carrier-only" channels. Computer interface capability, 24-hour clock/timer, and multiplexed output are among the features provided that allow the sophisticated listener tremendous expansion possibilities. An optional NTSC video will also permit monitoring of TV signals. The FRG-9600 is ready for operation from either 12 volts DC or 117 volts AC.

For further information, contact Yaesu Electronics Corporation, P. O. Box 49, Paramount, California 90723.

Circle #307 on Reader Service Card.

flea market

RATES Noncommercial ads 10¢ per word; commercial ads 60¢ per word **both payable in advance**. No cash discounts or agency commissions allowed.

HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing) on a space available basis only. Repeat insertions of hamfest ads pay the non-commercial rate.

COPY No special layout or arrangements available. Material should be typewritten or clearly printed (**not** all capitals) and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** cannot check each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue.

DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

MILITARY COMMUNICATIONS: CPRC-26 Manpack Radio (described in March 1985 Ham Radio), transceives 46-54 MHz FM, with battery box, antenna, crystal, handset: $22.50 apiece, $42.50/pair, good condition. R-390A Receiver, .5-30 MHz all modes, 4 mechanical filters, meters sealed (Government removed, operation unaffected): complete/checked $175, complete/unchecked $115, spare parts unit (80% complete, missing PTO/power supply) $65. Info SASE. CPRC-26 add $4/unit shipping. R-390A shipping charges collect. Baytronics, Dept. HR, Box 591, Sandusky, OH 44870. Phone (419) 627-0460 evenings.

CX7 REPAIRS. 415-549-9210.

LETTERING and text for QSL pasteup and label usage. Stamp brings circular. Labelcraft, PO Box 412, W. Sand Lake, NY 12196.

HOUSE READY FOR HAM. Beautiful mountain top all brick 4 bedroom, 3 full bath home and radio site on one acre landscaped, private, with huge inground pool. Located 55 miles from Washington, DC and 60 miles from Baltimore, easy commute by Amtrack or van pools. Extremely energy efficient brick home, full basement and other extra rooms, radio room. All cable, poles, antennas, fire alarm system and intercom system stay. Must see to fully appreciate. Call or write to receive free written details of all features, $250,000. (301) 371-7521. John Murray, Box 290, Jefferson, Maryland 21755.

TENNATEST — Antenna noise bridge — out-performs others, accurate, costs less, satisfaction guaranteed, $41.00. Send stamp for details. W8UUR, 1025 Wildwood Road, Quincy, MI 49062.

CUSTOM MADE embroidered patches. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept. 301, 4202 N. Drake, Chicago, IL 60618.

RECLAIM SILVER from electronics scrap! Write RALTEC, 25884F Highland, Cleveland, OH 44143.

HF QRP KIT, 5 watts output on 40, 30, 20 meters $39.95 plus $2.00 shipping includes one crystal, your choice of 7.035, 7.125, 10.108, 10.125 or 14.035 MHz. Write for full data and schematic. RF Kit Co., Box 27127, Seattle, WA 98125.

CUSTOM EMBROIDERED Emblems — Cloisonne' enameled pins, your design, low minimum, excellent quality, free booklet. A.T. Patch Co., Dept. 65, Littleton, NH 03561. (603) 444-3423.

AN/PRC-25 with accessories; handset, 2 antennas, canvas and battery. Working, $200. PRC-74 accessories; battery boxes, CY-6121 or CY-6314, new condition, $30 ea. PRC-74B manuals; depot $18 or operators $10. Looking for a PRC-74 transceiver in any condition. Stroh, N6JSD, 21930 Valerio St., No. 11, Canoga Park, CA 91303.

PLASTIC PACKAGE Sealing Tape: 4- standard 165' tan rolls, 1-3/4" $6; 1-1/2" $5. One roll 3" clear for sealing or label protection $2.50. All postpaid and satisfaction guaranteed. Lisaius, 116 Orton Rd., W. Caldwell, NJ 07006. (201) 226-7943.

WANTED 512-F1 P.S. Coffins; sell 50 year collection tubes; send list of requirements for price quotes. M. Levy, 4141 Krupp Drive, El Paso, TX 79902.

FOR SALE: Heath HW-101 $175.00. Robot 400 $225.00. Heath 2K tuner $85.00. TH5DX $175.00. KC0PT (913) 625-7141.

FOR SALE: *Hallicrafters SX-100 communications receiver.* Antique collectors item. Good condition. $139 or best offer. Gerry Nemetz, W4NEX, 8202 Beechwood Drive, Lynchburg, VA 24502. (804) 239-7789.

FOR SALE: Yaesu FV 301; FP 301 speaker; FT 301S transceiver; FL 110 linear amplifier; dynamic mike; Drake matching network MN-4C; DSI 3550 frequency counter. $750 takes the package. Send money order. *Milt Kappstatter, 39* Glenview Drive, Springfield, NJ 07081.

PRACTICAL APPLICATIONS mathematics system on disk for C-64. Will calculate geometric shapes, linear regression, linear interpolation, and base change. A must for calculating mechanical aspects of tower and antenna construction. $29.95. Microenergy Systems, PO Box 5291, High Point, NC 27262. (919) 885-0433.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

CHASSIS and cabinet kits. SASE K3IWK.

IMRA, International Mission Radio Association, helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, NY 10538.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 09224 (213) 774-1255.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

JENNINGS UCSL-1000 vacuum variable capacitors, 10 thru 1000 MMFD at 5000V with gear drive train and mounting bracket. Perfect for that linear amplifier or antenna tuner, brand new $59.50, plus $3.00 UPS. Lab calibrated HP 608C signal generators 10 MHz thru 480 MHz. Perfect for HF/VHF/UHF repair $345.00. HP614A covers 900 MHz thru 2100 MHz, excellent microwave signal generator $345.00, order today, have good stock. M/C, Visa, or check. Phone Bill Slep 704-524-7519, Slep Electronics Company, Highway 441, Otto, NC 28763.

KT5B Antenna 160M-10M no traps! $59.95. Weather boot kit $8.95. Open-wire, roller-inductor, antenna accessories and much more. Kilo-Tec, PO Box 1001, Oak View, CA 93022. Tel: 805-646-9645.

CABLE TV EQUIPMENT. Jerrold, Hamlin, Zenith — many others. Factory Units/lowest dealer prices. Complete illustrated catalog $2.00. Pacific Cable Co., Inc., 7325½ Reseda Blvd., Dept. 1003, Reseda, CA 91335. (818) 716-5914.

BROWN BROTHERS PADDLE — 500 No address QSL cards. For swap or offer. Kenneth Hand, WB2EUF, PO Box 708, East Hampton, NY 11937.

RADIO OFFICER: FCC and USCG licenses, and 6 months endorsement required for seagoing employment. Excellent money and fringes for technically qualified person. Send resume and copies of licenses to: Box HR2413, 810 Seventh Avenue, New York, NY 10019.

RETAIL STORE being liquidated by mail auction. Parts, tools, test equipment, home supplies, SASE to ESI, Box 328, Mesa, AZ 85201.

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England

Holland Radio
143 Greenway
Greenside, Johannesburg
Republic of South Africa

Coming Events

ACTIVITIES

"Places to go..."

MICHIGAN: The "85" U.P. Hamfest, July 27 and 28, St. Francis de Sales School, Manistique. Friday evening Fish Fry, set and eyeball for early arrivals. Saturday 6 AM to 5 PM. Banquet 6:30 PM. Sunday 8 AM to 2 PM. Registration $3.50. Free baby sitting. Table space $3.00 per 4' table. For more information: Debbie Barton, WD8IBT, 509 Range St., Manistique, MI 49854. (906) 341-5694 after 3 PM.

SOUTH CAROLINA: Charleston Hamfest, sponsored by the Charleston Amateur Radio Society, July 13 and 14 at the Omar Shrine Temple, East Bay Street. 8:30 to 4:00 Saturday, 9:00 to 4:00 Sunday. General admission $5.00 includes admission to Hospitality Room, 7:30 PM to 11 PM Saturday. 12 and under free. FCC exams Saturday. Buffet available both days. Flea market tables $5. Commercial booths $40. Talk in on 146.19/79. For information contact: Hamfest Committee, PO Box 70341, Charleston, SC 29405 or (808) 747-2324 or 554-8058.

IOWA: The Des Moines Radio Amateur Association will hold an Electronic Fair, Airport Hilton Inn, Des Moines, July 20 and 21. The Electronic Fair combines the Iowa ARRL Convention with what was the Des Moines Hamfest. General public admission fee $2.00. Flea market parking $3.00. Saturday evening banquet $15.00 per person. Featured banquet speaker Nick Johnson. There will be seminars both days for computer and satellite enthusiasts, Amateur Radio operators and spouses. For further information: Des Moines Radio Amateur Association, PO Box 88, Des Moines, IA 50301.

MINNESOTA: The St. Cloud Amateur Radio Club Hamfest, August 11, Sauk Rapids Municipal Park, off MN Hwy 15 (Benton Drive). Displays, demonstrations and trades. Ticket donation $3. Extra ticket $2. Snack counter. Talk in 34/94 primary; 615/015 secondary. Contact: SCARC, Box 141, St. Cloud, MN 56302.

KENTUCKY: The Central Kentucky ARRL Hamfest, sponsored by the Bluegrass ARS, Sunday, August 11, 8 AM to 5 PM, Scott County HS, Longlick Road and US 25, Georgetown. Tech forums, license exams, awards and exhibits. Air conditioned facilities. Free outdoor flea market space. Tickets $3.50 advance and $4.00 at gate. Talk in on 76/16. For information or tickets SASE to Scott Hackney, KI4LE, 629 Craig Lane, Georgetown, KY 40324.

MISSOURI: The 23rd annual Zero-Beaters ARC Hamfest, July 21, 9 AM to 3 PM, Washington, MO Fairgrounds. Free admission. Free flea market area. Limited rental spaces under pavillion. Advance reservation advised. FCC exams, cake walk, candy scramble, traders tow, refreshments and food available. Talk in on 147.24-84, 146.52. For information: Zero-Beaters ARC, Box 24, Dutzow, MO 63342.

NORTH CAROLINA: The Western Carolina Amateur Radio Society's annual Hamfest, July 27 and 28 at Buncombe County Fireman's Training Center, Asheville. *Admission $4* at gate, $3.50 advance. Free parking, camping (no hookups), forums, and VEC exams will be available. Outside flea market sites — bring own table. Talk in on 16/76 and 31/91. For advance tickets contact: Marvin Soloman, KI4EA, 14 Carjen Avenue, Asheville, NC 28804. All other inquiries to: Earl Elliott, KI4UO, 17 Emory Road, Asheville, NC 28806.

PENNSYLVANIA: The Foothills ARC's 17th annual Greensburg Hamfest, Nevin Arena, Sunday, July 28. Tickets $2.00 or 3/$5.00. Indoor tables $5.00. Tailgating $2.00. Refreshments. Mobile check in on 147.78/18. For further information, registration or tables: F.A.C.R., PO Box 236, Greensburg, PA 15601 or contact WB3KJH.

INDIANA: The WA9SNT Amateur Radio Club will hold its annual Swapfest, August 4, ITT Technical Institute, 9511 Angola Court, Indianapolis. 8 AM to 4 PM. Admission $2.00. Students $1.00. Flea market space $1.00 addit. Talk in on 146.94 and 3910 ± . For information: Dave Johnston, K9HDQ, ITT Technical Institute, 9511 Angola Court, Indianapolis, IN 46268. (317) 875-8640.

FLORIDA: The 12th annual Greater Jacksonville Hamfest, August 3 and 4, Jacksonville Civic Auditorium, Saturday 8 to 5 PM; Sunday 9 to 3 PM. Registration $4.00. Children under 16 admitted free. Reserved swap tables $9.00/1 day; $15.00/both days. FCC exams for all grades Saturday at 1 PM on a walk-in basis. For information and registration: SASE to Jacksonville Hamfest Assn., PO Box 23134, Jacksonville, FL 32241.

MICHIGAN: The Straits Area ARC is having its 12th annual Swap 'N Shop, July 20, Emmet County Fairgrounds, Petoskey. 9 AM to 2 PM. General admission $2.50. Single table $3.00. Refreshments available. Free parking Friday night for self-contained RV's. Petoskey State Park nearby. Come and bring the family. Talk in on 07/67 and 52. For further information: Joe Warden, WD8MJB, PO Box 444, Conway, MI (616) 347-8693. Please SASE.

1985 BLOSSOMLAND BLAST, Sunday, October 6, 1985. Write "BLAST", PO Box 175, St. Joseph, MI 49085.

WISCONSIN: The Oshkosh Amateur Radio Club in conjunction with the S.O.L.A.R. Assn. will host EAA hams for the 1985 convention, July 26 - August 2. Stop and rest, charge your batteries, leave messages, etc at the EAA Ham Shack located at the north end of the commercial exhibit area. Look for the red and white ARRL flag. On Saturday, July 27 at 3 PM, there will be a gathering for all EAA hams hosted by the Oshkosh ARC. We'll be serving bratwurst, burgers and refreshments free of charge. Bring your wives and kids. You're in for a treat! For further info: Forest Schafer, WD9IWL, 417 Willow St., Omro, WI 54963.

WEST VIRGINIA: Wheeling Hamfest and Computer Fair, Wheeling Park, Sunday, July 21. Dealers welcome. Flea market, ARRL, AMSAT, SWOT, SMIRK booths. Family activities available at Park. Admission $3.00. To reserve space contact Jay Paulovicks, KD8GL, RD 3, Box 238, Wheeling, WV 26003. (304) 232-6796 or TSRAC, Box 240, RD 1, Adena, OH 43901 (614) 546-3930.

CQ CONTEST: VHF'ers please note! The first annual CQ World Wide VHF WPX Contest is July 20-22, 50 thru 1296 MHZ. For details, logsheets, etc., write to SCORE, PO Box 1161, Denville, NJ 07834 or to CQ Magazine. We need your entry to make this a success.

WISCONSIN: The South Milwaukee Amateur Radio Club's annual Swapfest, Saturday, July 13, American Legion Post #434, 9327 South Shepard Avenue, Oak Creek. Activities start about 7 AM and will run through 4 PM. Parking, picnic area, food and refreshments available. Free overnight camping. Admission $3.00 and includes a free beverage. The Milwaukee Volunteer Core Group will conduct Amateur Radio exams during the day. For more details and a map: South Milwaukee ARC, PO Box 102, South Milwaukee, WI 53172-0102.

NEW JERSEY: (Augusta) July 20: The Sussex County ARC will sponsor SCARC '85 at the Sussex County Fairground, Plains Rd., off Rt. 206. Doors open 8 AM. Registration $2.00. Indoor tables $7.00 each. Tailgate space $5.00. Food and refreshments. Free parking. Talk in on 147.90/30 and 146.52. For further information: Donald R. Stickle, K2OX, Weldon Rd., RD #4, Lake Hopatcong, NJ 07849. (201) 663-0677.

CALIFORNIA: The first International Youth Tele-congress will convene in Santa Cruz, July 19 to 23. The goal is to link young people around the world via Amateur Radio and computer bulletin board networks. For information: Redwood Youth Foundation, 5300 Glen Haven Road, Soquel, CA 95073. (408) 476-2905 or (408) 662-0300. WA6KFA.

WEST VIRGINIA: The 7th annual TSRAC Wheeling Hamfest and Computer Fair, Sunday, July 21, Wheeling Park, 9 AM to 4 PM. Easy access, dealers, exhibits all under roof. 5 acres Flea Market. Refreshments, free parking, family park activities. Admission $3.00. ARRL, AMSAT, SWOT, SMIRK, etc. For information and map: Jay Paulovicks, KD8GL, RD 3, Box 238, Wheeling, WV 26003. (304) 232-6796 or TSRAC, Box 240, RD 1, Adena, OH 43901 (614) 546-3930.

NORTH CAROLINA: The 13th annual Mid-Summer Swapfest, sponsored by the Cary ARC, Saturday, July 20, 9 AM to 3 PM, Lion's Club Shelter, Cary. Talk in on 80-30, 146.28/.88, 30-2, 147.75/.15, 2-0, 146.52/.52. For information: Cary ARC, PO Box 53, Cary, NC 27511.

BRITISH COLUMBIA: Okanagan International Hamfest, July 27 and 28, Oliver Centennial Park, Oliver. Registration July 27 at 9 AM. Activities Saturday, July 27 at 1 PM and Sunday July 28 at 2:30 PM. Saturday potluck supper. Talk in on 146.34/94 OKN Rpt 76/76. For further info: Lota Harvey, VE7DKL, 584 Heather Rd., Penticton, BC V2A 1W8. (604) 492-5768.

BRITISH COLUMBIA: Maple Ridge Hamfest, July 13 and 14, St. Patricks Center, 22589 — 121 Avenue, Maple Ridge. Admission: Hams $5.00; non-hams $2.00. Food, swap & shop, commercial displays, bunny hunt and family activities. Close to shopping and swimming. Camper space, no hookups. Talk in on 3.758 MHz 146/20/80 and 146.34/94. For information and a 20% pre-registration discount: Maple Ridge ARC, Box 292 Maple Ridge, BC, Canada V2X 7G2.

MARYLAND: The Baltimore Radio Amateur Television Society's famous BRATS Maryland Hamfest and Computerfest, Sunday, July 28, Howard County Fairgrounds, Rt. 144, near I-70, West Friendship. Over 175 tables all indoors. Tailgating $3.00 per space. RV hookups available on grounds. Nearby motels. Free walk-in VE exams. For further information/table reservations: Mayer Zimmerman, W3GXK, BRATS, PO Box 5915, Baltimore, MD 21208.

NEW YORK: The Mt. Beacon ARC Hamfest, Saturday, July 20, Arlington Senior High School, Poughkeepsie/Lagrange. Tickets $2.00. Non-hams and children admitted free. Tailgate space $3.00. Tables $4.00. Doors open 8 AM. Talk in on 146.37/97 and 146.52. For information: Julius Jones, W2IHY, RR2, Vanessa Ln, Staatsburg, NY 12580. (914) 889-4933.

COLORADO: Amateur Radio Motorcycle Club Rocky Mountain Roundup III will be held somewhere west of Denver. Riding radio operators check the ARMC net Thursday nights, 0300 UTC, 7237.5 kHz. Send business SASE to AG0N, Gary McDuffie, Rt. 1, Box 464, Bayard, NE 69334 and ask for net information.

MASSACHUSETTS: The first ARRL Heavy Hitters Hamfest, July 20 and 21, Topsfield Fairgrounds, US 1, Topsfield. 9 to 4 both days. Giant flea market, ARRL, PACKET, AMSAT, ATV and more. License exams held at nearby school. For reservations send completed 610 form and $4.00 check payable to ARRL/VEC, copy of current license and SASE for confirmation to: Topsfield Exams, c/o PO Box 71, Hanover, MA 02339 by June 21. Sorry no Novice exams. Free camping Saturday night for tents and self-contained RV's. Nearby hotels. Advance tickets $3.00. $4.00 at door. Non-ham spouses and kids admitted free. Talk in on 146.64 and 147.285 repeaters. For information: Russ Corkum, WA1TTV, 21 Thorndike Street, Arlington, MA 02174.

OPERATING EVENTS

"Things to do..."

The Eastern Michigan ARC will commemorate the annual Port Huron to Mackinac island Yacht Race, July 20 and 21 from 1400 Z to 0200 Z both days. Listen for K8EPV. For a certificate send QSL with legal-size SASE to K8EPV (C.B.A.) or 654 Georgia, Marysville, MI 48040.

A direct Trans-Atlantic QSO on 2 meters — can it be done? The attempt is being organized by the West Kent Amateur Radio Society and will take place between August 19 and 30. Arrangements for skeds (high power stations only please) can be made by contacting Dave Green, G4OTV, 13 Culverden Down, Tunbridge Wells, Kent, TN4 9SB, England. Tel: 892-28275.

Vaynesboro, Virginia Parks and Recreations Department and the Valley Amateur Radio Association will operate special event station KI4BR in Ridgeview Park in celebration of "Summer Extravaganza". 1700 hours C.T.U. on Saturday and Sunday, July 13 and 14. A "First Edition Certificate" will acknowledge QSO and receipt of QSL. SASE to KI4BR, PO Box 565, Waynesboro, VA 22980.

The Southern Michigan Amateur Radio Society will operate W8DF/8 during the 7th World Hot-Air Balloon Championship, July 13 to 21, Kellogg Regional Airport, Battle Creek, Michigan. Phone: center of General 80-10 meters and CW in Novice bands. For a special QSL SASE to PO Box 934, Battle Creek, MI 49016.

Riding Radio Operators — Amateur Radio Motorcycle Club Net meets every Thursday night at 0300 UTC at 3888 kHz standard time and 7237.5 kHz daylight saving time. An eastern USA group meets one hour earlier at 3888 kHz year-round. Send business SASE to AG0N, Gary McDuffie, Rt. 1, Box 464, Bayard, NE 69334 and ask for net information.

July 7-13: The Cherryland ARC will operate special events station KA8QVH to commemorate the National Cherry Festival, Traverse City, MI. 1100Z July 7 thru 0200Z July 13 daily. Send large SASE with QSL to Ed Irwin, 346 Peninsula Trail, Traverse City, MI 49684 for an attractive certificate.

July 27: CARS Third annual SPACE DAY special event station will be on the air 0000 GMT July 27 to 1900 GMT July 28. For a certificate send QSL and $1.00 postage to CARS, PO Box 512, Jackson, MI 49204.

July 19, 20: The Indian Hills Community College ARC will operate special events station WA0IUQ during the 1985 Ottumwa Hot-Air Balloon Races. 2200 GMT to 0400 GMT on SSB only each day. For a commemorative QSL send your QSL with SASE to WA0IUO at Callbook address.

WIA 75 AWARD: March 1 to December 31, 1985: The world's first and oldest radio society, Wireless Institute of Australia is celebrating its 75th anniversary. To qualify for the WIA award, Australian Amateurs and SWL's must log 75 members of WIA: Overseas Amateurs and SWL's contact station VK75A; contact Amateur already qualified for WIA 75 award; contact 75 WIA members and log membership numbers. Send info plus $2 for certificate and s&h to: WIA 75 Award Manager, Wireless Inst. of Australia, 412 Brunswick St., Fitzroy 3065, Victoria, Australia.

July 25: The Kauai Amateur Radio Club is planning an expedition to Hawaii's 5th county, Kalawao County on the island of Molokai. Callsign KH6F on July 25, 26, 27 and 28. 80, 40, 20, 15 10 and 2 meters. SSB, FM, CW. Send QSL and SASE or IRC to KH6F, PO Box 675, Koloa, HI 96756.

THE GUERRI REPORT

Ernie Guerri W6MGI

maybe — and that's final!

The earliest developers of logical concepts were essentially philosophers who accommodated the notion that not every question had an answer that was clearly black or white. "It all depends . . ." was a perfectly valid answer to some questions.

During the Renaissance there was a significant effort to minimize indeterminate answers to clear-cut questions, mostly as an accommodation to the growing influence of scientists, for whom things either "were" or "weren't." This trend has continued to the present, with a growing gap between the "certainty" of scientific dogma and the "softness" of philosophic postulation. The birth of the electronic computer age in the middle of this century gave the final imprimatur to determinate logic; everything was either one or zero, and every entity in the universe could be logically diminished into finite elements which either "were" or "weren't."

Two important developments in modern science are making things less certain than they seem to have been between 1500 and 1950. First, the observation of over 200 "basic" atomic characteristics has many theoretical physicists contemplating the concept that the physical universe may extend indefinitely in both the micro and macro directions. Second, the desire to implement artificial intelligence in robot devices has given renewed credence to the philosophers' acceptance of "maybe" as a valid answer. This new class of indeterminate logic, or "fuzzy logic" as some call it, is characterized by the general concept that the consequences of events are not necessarily certain, nor need they be — and that some inferences which have no obvious mathematical value can improve the precision of a logical decision. A clearer picture of these concepts may be had from a graphic representation (**fig. 1**).

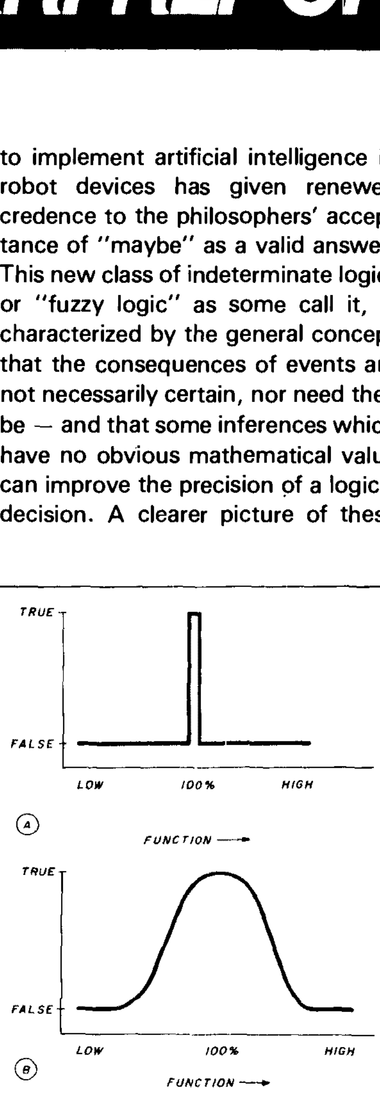

fig. 1. (A) Determinate logic, in which "OK" has a narrow range and (B) indeterminate logic, in which "OK" is not so clear.

The bottom line for all of this is actually a reliability issue. If devices can be designed to operate with some form of judgment that allows them to weigh external influences, or assess their own degradation, then perhaps these devices will function longer, or with less error, than devices that will only "run" or "break." In the indeterminate case we have the option of continuing the function if things are "true enough" to be "mostly OK." Deciding that you have "enough" air in a "flat" tire to make it to the next gas station is an example. This "new" class of logic will be interesting to observe as it is implemented in artificially intelligent devices.

FM broadcasters set spectrum standards

In April, 1984, the FCC deregulated the way in which commerical FM broadcasters (88-108 MHz) can use their assigned 200 kHz spectrum slot. A stereo audio broadcast uses 106 kHz (±53 kHz) of the assigned slot, and the FCC has said that the broadcasters are free to have as many subsidiary communications authorization (SCA) channels in their slots as they like. The broadcasters have been quick to recognize that the unused 94 kHz (±47 kHz) is valuable spectrum that reaches into every home and business in their service area. Among the applications that

they have implemented in the past year are 9600 baud videotext for stock and commodity businesses. This data rate allows a full screen update every 6 seconds. In addition, the data stream can be coded to provide different data to different subscribers.

Additional uses now include traffic alerts sent to properly equipped car radios, full color graphics transmission (at about 10 seconds per screen), radio teleshopping, and perhaps the most financially rewarding application, personal paging. This last use is significant because of the excellent coverage most FM stations get by virtue of their high power and advantageous antenna siting. The applications are limited only by how many subcarriers and modulation techniques the broadcasters can stuff into the allocated space, yet still maintain the requisite quality for each transmission mode.

For Amateurs who complain that we don't have enough space to do our thing, the FM broadcasters are setting an example of efficient spectrum utilization that bears watching.

high speed health hazards

The fabrication of very high speed digital and microwave semiconductor devices requires the use of many exotic materials. Gallium and indium arsenides, phosphorus, and cyanide recovery agents are among the many materials in regular use. Unfortunately, these materials that are so beneficial to technology are deadly to most living things. The cities surrounding California's Silicon Valley are now observing disturbing levels of many of these undesirable materials in local water supplies. The omens are not good. The entire electronic industry must view this matter with urgent alarm, lest the plants — which have caused our industry to flower — die from their own droppings.

ham radio

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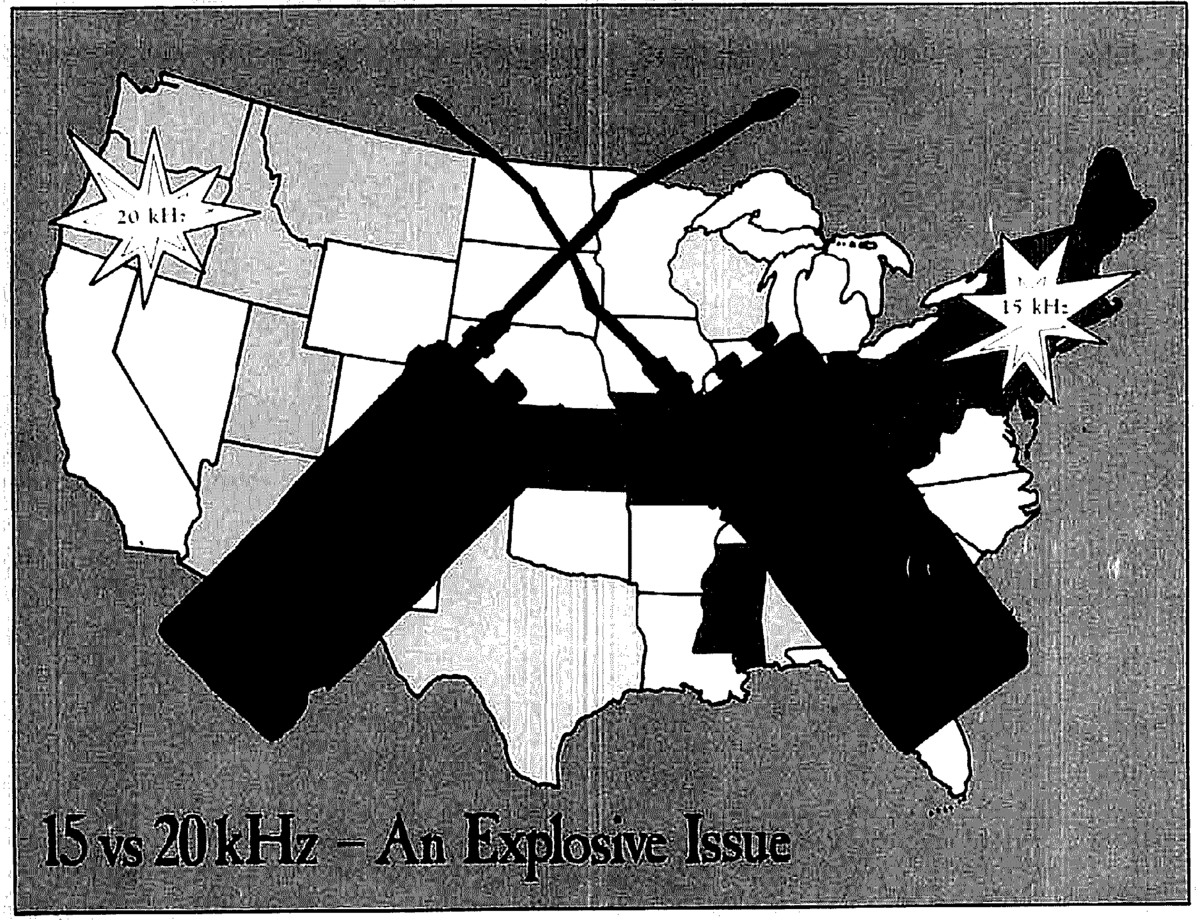

hr

focus on communications technology

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ham radio

magazine

AUGUST 1985

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contents

presstop de W9JUV

STILL ANOTHER THREAT AGAINST 220 MHZ HAS BEEN MOUNTED by an oil exploration-related firm in Illinois. LAOAD Radio and Microwave Communications Consultants petitioned the FCC to allocate 216-220 MHz for use as 1033 ASCB data and voice channels, AND provide a 350 kHz segment of 220-222 MHz to be shared with Amateur Radio. Under terms of the proposal, LAOAD would advise the ARRL of pending operations, and the League would then alert Amateurs.
Little, If Any, Interference To Amateurs Would Result, the petitioner (W9GT, whose Norwegian call LAØAD inspired his company name) claims, as operations would be very short-lived and typically well removed from populated areas. However, 220 MHz users and others are very concerned that any such incursion could set a dangerous precedent for the future.
The Original Comment Period On RM-4983 Had Closed by presstime, but there has been considerable pressure to reopen it. Though the recent upsurge in interest in providing Novices with 220 MHz privileges (see below) certainly tends to diminish or at least postpone threats such as this one, it is still quite serious and should be challenged.

ARRL'S PETITION TO EXPAND NOVICE PRIVILEGES HAS RECEIVED its FCC Rule Making number, RM-5038, along with several other related petitions. In brief, the League proposal would give Novices limited 10-meter SSB and data privileges plus full privileges (with reduced power) on 220 and 1246-1260—see July Presstop for specific details. Somewhat similar petitions filed by KC5OQ include giving Techs and Novices ASCII (RM-5022)and phone (RM-5032) on part of 10 meters, some 30 meter privileges (RM-5024), and 220 MHz phone (RM-5025). The Comment period closed for KC5OQ's petitions July 11; a date hadn't been set for ARRL's at presstime, but its cutoff date should fall sometime in late July.

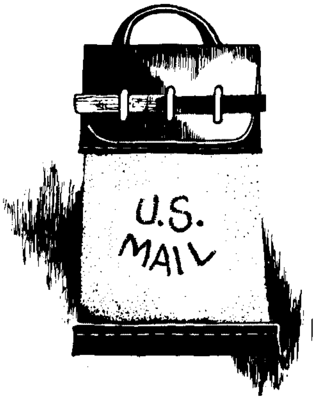
BROADCAST STATIONS MAY RETRANSMIT AMATEUR TRANSMISSIONS or use what they hear on the Amateur bands on the air, but any direct involvement between Amateurs and a broadcast station is expressly forbidden. In its June 7 Report and Order on BC Docket 79-47, the Commission agreed that the content of an Amateur transmission is not protected against reuse by others, but that to have, for example, an Amateur station at a broadcast studio to solicit traffic information during rush hours would be against the rules.

AMATEURS MAY BE ALMOST TWICE AS PRONE TO LEUKEMIA as the general population, a study reported June 23 by the New York Times News Service suggests. Underscoring ham radio Editor K2RR's July editorial, Washington state epidemiologist Dr. Samuel Milham, Jr. found the death rate from various forms of leukemia for 1691 California and Washington Amateurs who died between 1971 and 1983 was just about twice as high as would normally be expected. The increase was in myeloid and unspecified forms of leukemia; lymphatic and monocytic forms of the disease had no higher incidence among Amateurs than among others.

COMMENT DEADLINE FOR FCC'S NATIONAL REPEATER COORDINATION proposal (PR Docket 85-22) has been extended to August 15 in response to an ARRL request. The new Reply Comment date is September 30. Comments received thus far by the FCC have almost all been endorsements of the concept of national coordination, with few specific ideas on how to make it work.

VEC MAINTENANCE OF AMATEUR QUESTION POOLS WAS PROPOSED by the FCC in a Notice of Proposed Rule Making issued June 12. VEC-developed questions would have to follow the Commission's syllabus, and each pool would have to include at least 10 times the number of questions asked in that particular exam. At the same time the FCC also proposed moving up the date at which VECs may begin preparing their own exams. Comments on PR Docket 85-196 are due at the Commission by August 30, and Reply Comments by September 30.
One Regional VEC Has Already Dropped Out Of The Program, and another has said it plans to when the ARRL gets fully up to speed. Still another VEC has asked his Senator to look into the fee structure, as he feels his club isn't getting enough money for its efforts. For comparative purposes, NABER's fee for administering an exam for certification under its commercial examination program is $38!
The FCC's August VEC Meeting In Gettysburg Had Relatively Few signed up at presstime, despite a very promising program. Planned topics include improving speed of service, decreasing paperwork errors, and improving the integrity—real and perceived—of the program. Attendees will also meet the FCC people they've been dealing with and see the FCC's licensing facilities and procedures in operation.
Volunteer Examiners Are Now Averaging About 4000 Exam Elements Monthly, compared to a high of about 2750 exams a month when the FCC was still giving exams. However, the U.S. Amateur population is still well under its March, 1983, all-time high of 414,973; latest (May 30) figures show 410,846 individual FCC-licensed Amateurs.

U.S. AMATEURS MAY FINALLY USE OSCAR 10'S MODE "L" TRANSPONDER, under an FCC STA issued June 12. The Special Temporary Authorization permits uplink transmissions to OSCAR 10 from 1269.05 to 1269.85 MHz; an editorial error in the initial FCC release limited the STA to Extras only, but that was later corrected to include Technicians and above.



comments

can we talk?

Dear HR:

In the May, 1985, editorial it was noted that we Amateurs can't really communicate. I find this statement to be offensive, but in many respects accurate....

I find it offensive because I think it comes from a narrow perspective. It seems the persons making such statements see only one aspect of our hobby. I would agree that at times we really don't communicate much of anything interesting or important. The hard work and learning that made such contacts possible can't be judged on this evidence alone.

The major reason for this lack of ability to communicate clearly and effectively is a failure of our society and its education system, not the fault of Amateur Radio. We in Amateur Radio are only a reflection of the society which makes our hobby possible. Everywhere one looks these days, one can find someone writing about the deplorable state of our schools. We have students who can't read or write properly receiving diplomas and college degrees. To correct this sad state of affairs, we as a society will have to change our approach to education and discipline....

One last issue before I close. I don't think that many people, when they judge our ability to communicate, listen to 2 meters FM. This activity is one really bright spot in Amateur Radio when it comes to communication. Here you find truly local activity being carried out. In the case of my area, Southern New England, we have many fine activities taking place on repeaters and on simplex. We have a computer net on the W1XJ repeater that meets on Mondays at 8:30 PM, local time, where hams communicate ideas and educational information concerning this aspect of our hobby. We have several MSO-type RTTY bulletin board systems up and running, distributing Amateur information of all kinds. We now have active packet radio systems in operation with all sorts of bulletin boards, message systems, and other features that promote real communication.

Kenneth E. Stringham, Jr., AE1X
Attleboro, Massachusetts

service — not hobby

Dear HR:

Congratulations on an outstanding May issuel

It was especially interesting to read the editorial, "The Readers Speak," (page 4) regarding the problems and solutions before us. Let me call to your attention one of the problems that seems to have escaped your attention. It is one of perception, and it is reflected by you in your editorial by the use of the word "hobby" no fewer than five times. Of course, you are referring to the Amateur Radio "Service." It is, you know, a federally regulated "service" and not a "hobby."

Stamp collecting is a hobby. Model building is a hobby. Woodcraft is a hobby. Amateur Radio is a federally regulated *service* — even though it is perceived as a hobby by all too many hams and would-be hams.

Typically, hobbies don't require formal exams and licenses and involvement with federal regualtions. One has only to read Part 97 to see that the purpose and intent of Amateur Radio was not to establish a hobby, but rather a federally regulated communications service for the public interest, convenience, and necessity. Indeed, too many of us are overly involved in contesting and card-collecting — the hobby aspects of ham radio. But if it's presented to us as a hobby, why not?

Our organization, the Wireless Institute of New Orleans, was established by a group of dedicated hams to preserve the original principles of Amateur Radio and to promote the state of communications art. We continue to observe the degradation of ham radio into what appears to be an expanded version of the Citizen's Band. But what the hell? It's just another hobby, isn't it?

A.J. ("Buddy") Massa, W5VSR
New Orleans, Louisiana

matching dipoles

Dear HR:

Even though George A. Wilson, Jr., W1OLP, in "Matching Dipole Antennas," (May, 1984, page 129) made at least 24 separate references to GDO (Grid Dip Oscillators) and Grid Dipping, someone is certain to try substituting a solid-state dipper, (such as the Heathkit HD-1250 or one of several factory assembled versions) when exciting the RF Bridge discussed in the article. In fact, with the solid-state dipper far more prevalent today than the old vacuum tube grid dip oscillator (and interchangeable in most applications), no doubt a large number of hams who build the RF bridge will end up frustrated and with no discernible "dip."

While the solid-state dippers can be used to determine resonance, per the first part of George's article, it is not likely to provide enough excitation to obtain a reading with the RF bridge unless overcoupled, with sensitivity set at maximum, and with an extremely sensitive μA meter used as the detector. Even a 50 μA meter will probably not allow a discernible "dip" to be obtained!

A rough idea of a dipper's suitability can be obtained by connecting a germanium diode and a small 2 to 3-turn link in series across the μA meter's terminals. Coupling the link to the dipper's coil should easily produce a full-scale reading. If it does not, the dipper cannot be used to excite the RF bridge.

Robert G. Wheaton, W5XW
San Antonio, Texas

FM repeater separation — 20 kHz *Yes*, 15 kHz *No*

Proving the point through VHF FM receiver selectivity measurements

Amateur use of the 2-meter **(144-148 MHz)** band is now under nationwide scrutiny in an effort to determine whether the channel spacing for FM sections of the band should be set at 15 kHz or 20 kHz. The original 30 kHz spacing was divided, as band use increased, into 15 kHz channels to allow more channels; this division led to increased adjacent channel interference in many areas, which in turn resulted in the current proposal to increase the channel spacing to 20 kHz.

Changing to the 20 kHz spacing will, of course, change the frequencies of some of the channels and change the overall number of repeater "pairs" in the band. Only the technical — not the political or emotional issues implicit in these changes — will be addressed in this article.

In trying to become better informed on the issue and thus establish a more substantial foundation for our decision in northern Colorado, we examined the nature of frequency modulation and its transmission and reception, and then made some measurements on several popular transceivers. We hope this information will be useful to other repeater groups and coordinators as they weigh this issue for themselves.

Our measurements were made to establish the actual performance levels of Amateur ("consumer") and professional ("commercial") receivers, with respect to adjacent channel rejection and variation of sensitivity with transmitter deviation setting.

frequency modulation

One factor that complicates any discussion of FM channel spacing is the varied levels of the understanding from one person to another of just how FM works. The following brief review may help to clarify the subject and shed some light on interpretation of our data.

In FM operation, the radio frequency output spectrum components vary as a function of the modulating (voice) signal amplitude. The resulting signal consists of a varying amplitude carrier and sideband pairs. (In narrow-band FM-only, the first sideband pair and carrier are significant in amplitude.) The amplitude of the carrier and sidebands is described by a mathematical term called a Bessel function of the first kind. The only thing we need to understand here is how much power is spread over how much spectrum, and what determines the signal (spectrum) width. Note that regardless of individual sideband or carrier amplitude, the *total* power of the FM signal is constant.

A simplified FM signal spectrum is illustrated in **fig. 1**. With no modulation applied, a single carrier term at a frequency f_c is visible. As the amplitude of the modulating signal is increased (from zero), a sideband pair displaced $\pm f_m$ from the carrier frequency appears. In this simplified version, we have assumed that a single-tone modulating signal (at frequency f_m) is used. Further increases in modulating signal amplitude cause additional sideband terms (pairs) to appear. At the same time, the amplitude of the carrier decreases. It is worthwhile reiterating that the total power of the FM signal is constant. This power distribution is a function of the modulation index β, which is defined as the ratio of frequency deviation (swing from carrier frequency) to modulating frequency (f_m). For small values of β, the bandwidth occupied by an FM signal is simply $2 \times f_m$. As β increases, more sidebands appear (separated f_m in frequency from each other). A natural further complication is that voice modulation can be considered to consist of many tones of varying

By Chris Kelly, WD5IBS, 1220 East Stuart #14, Fort Collins, Colorado 80525, and **Virgil Leenerts, WØINK**, 1007 W. 30th, Loveland, Colorado 80537

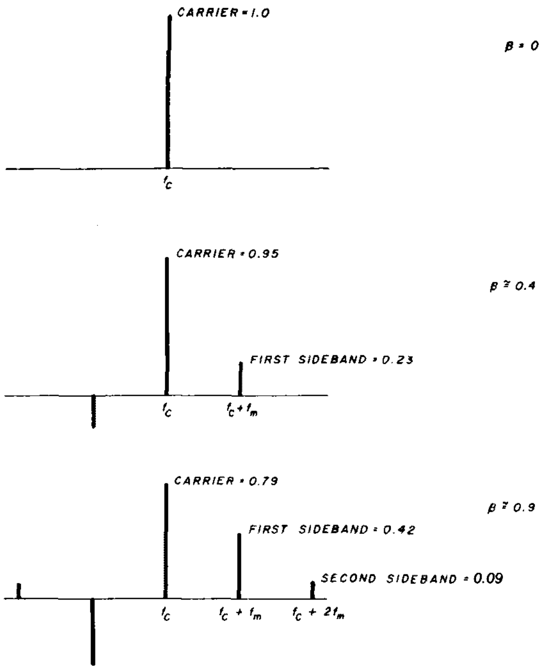


fig. 1. The spectrum distribution of an FM signal is a function of the modulation index β which in turn depends on the amplitude of the modulating (voice) signal. (*A*) the unmodulated carrier, (*B*) at a β of 0.4 a single sideband pair is evident with approximately 1/4 of the amplitude of the *un*modulated carrier, and (*C*) by increasing β to 0.9 a second sideband pair is also apparent. Notice that in both (*B*) and (*C*) the carrier power is reduced from its unmodulated value.

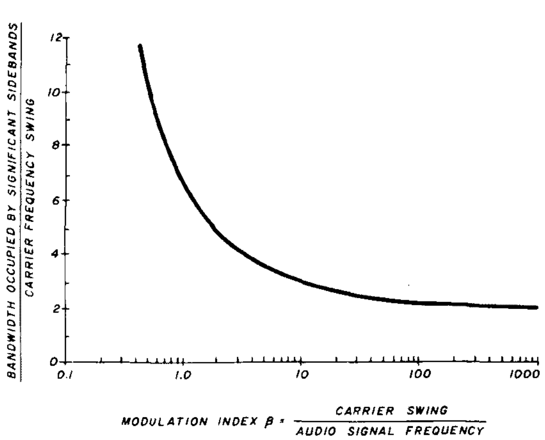


fig. 2. Bandwidth (normalized) versus modulation index β. (Adapted from C.E. Tibbs and G.G. Johnstone, *Frequency Modulation Engineering*, John Wiley & Sons, Inc., New York, as reproduced in *Reference Data for Radio Engineers*, Sixth Edition, Howard W. Sams, Inc., publisher.)

amplitudes. Consequently the total FM signal spectrum is quite complex.

For most VHF FM communications transceivers, this is 5 kHz deviation over 3 kHz maximum voice (modulation) frequency, or a β of 1.7 for high-pitched tones. Notice that lower deviation causes a lower modulation index. Using these figures, we find that 99.99 percent of the power in an FM signal will be contained in about 22 kHz of spectrum.[2] Depending on the assumed voice characteristics, this figure will change, and the older EIA specifications say that 99.99 percent of the power will occupy 19 kHz of spectrum.[3]

In the case of several FM signals, we do not have just narrow carriers that must be separated — we have finite bandwidth modulated signals occupying some spectrum.

For any given modulation frequency, we can decrease the modulation index, and thereby decrease the spectrum occupied, but not always in an exactly linear way. By increasing or decreasing the transmitter deviation control, the power ratios in the various sidebands will change, causing various effects on the radio channel and on the receiver.

effect of transmitter deviation on system performance

The Amateur 2-meter FM system is based on the commercial 5 kHz deviation FM system. System performance depends on the design and adjustment of the transmitters and the receivers used. However, design tradeoffs do exist.

Amateurs often discuss the effect of changing the deviation setting of ham transmitters, both in bandwidth and in effects on the receiver. We examined these two issues and made measurements of consumer gear and test equipment.

Figure 2 shows a curve of normalized significant bandwidth versus modulation index. Most Amateur transmitters adjusted for 5 kHz deviation will operate at a modulation index ranging between 3 and 6, depending on the operator's individual voice characteristics. The curve shows that in this range the curve begins to flatten, and that increasing deviation has less effect on bandwidth than at lower modulation indices. The "rules of thumb" used to roughly describe the bandwidth of FM signals involve a limited range where the slope of this curve can be considered constant. This is because as you decrease transmitter deviation, the modulation index for a given tone rises, changing the relative energy in each sideband.

Figure 3 illustrates the effect of the modulation index on the relative amplitude of FM sideband pairs. Consider the case of a 1 kHz tone, with the operator varying the deviation control on the transmitter. When the deviation control is at zero, all the RF power is contained in the unmodulated carrier. When the deviation rises to 1 kHz, the modulation index equals 1, and we

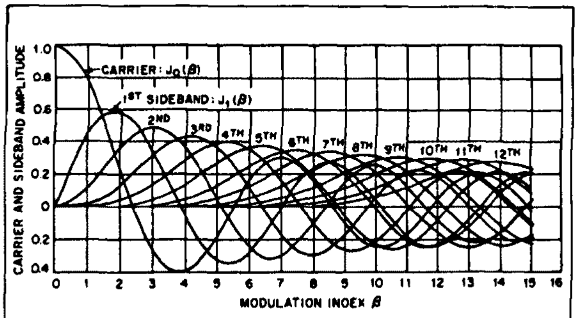


fig. 3. Plot of Bessel function of first kind as a function or argument β. (From P.F. Panter's *Modulation, Noise, and Spectral Analysis*, as reproduced in *Reference Data for Radio Engineers*, Sixth Edition, Howard W. Sams, Inc., publisher.)

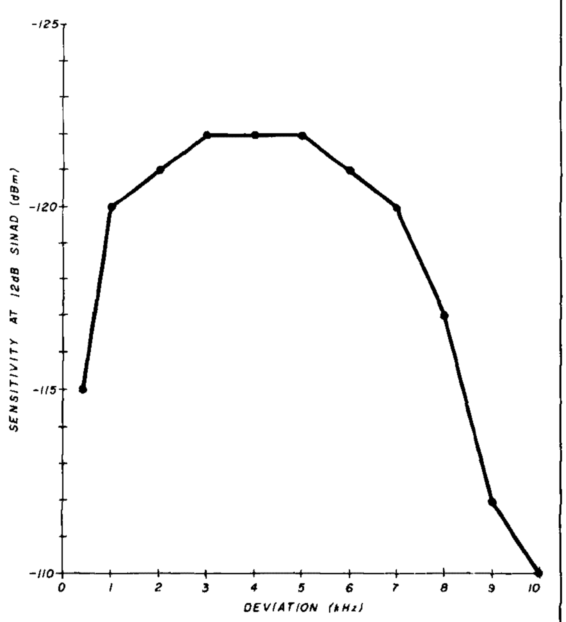


fig. 4. Effect of daviation on sensitivity for a popular Amateur 2-meter transceiver.

see a decrease in carrier power and increases in the first and second sidebands. In fact, there are increases in every sideband, but they are too small to show on this chart. At 1 kHz deviation, we see that the amplitude of the first sidebands has risen to about 0.44 times the original carrier level, and each sideband contains about 19 percent (0.44 squared) of the RF power. Now each of the second sidebands has about 1 percent (0.11 squared) of the RF power, and the carrier has only about 60 percent of the power.

As we raise the deviation to 5 kHz, the modulation index rises to 5 (5 kHz deviation/1 kHz modulation) and we can see that significant energy is now found in almost all sidebands up to the eighth. (Actually, there is energy present in other sidebands, but this chart cannot illustrate that.) The sidebands are spaced at intervals corresponding to the frequency of the modulating tone (1 kHz).

Note also how the modulation index varies with the modulating tone. Consider what would happen if we left the transmitter at 5 kHz deviation, but raised the modulating tone to 2000 Hz. The modulation index would drop to 2.5, and we would have to examine **fig. 3** at this new point to determine the relative amplitude of sidebands at the new index. Here, only the first five sidebands are noticeable — but remember, these sidebands are now 2 kHz apart. The bandwidth of the signal has increased, but it has not doubled.

It should be noted that this discussion of single-tone modulation is a very simplified version of what happens when voice is used to modulate the carrier. The voice is composed of many frequencies, and the composition changes with time. The components of the FM signal are many, and not just the sum of the voice frequencies. Consider a case of just two tones modulating the carrier. There will be carriers with amplitude of the Bessel function (J_0) at the deviation ratio of the first tone, the Bessel function (J_0) of the second tone, and sidebands having lines of all Bessel functions of f_1, f_2, $f_1 + f_2$, $f_1 - f_2$, $f_1 + 3f_2$, $3f_1 + f_2$, and so on.

If you now consider the complexity of the human voice, the problem of mathematically describing the bandwidth becomes unmanageable, at least for this author. For this reason the discussions here are limited to single-tone modulation.

The second aspect of performance affected by the deviation adjustment of the transmitter is how well the receiver is able to demodulate these signals. This is a very easily measured parameter. We checked the perormance of an Amateur receiver when receiving signals at different deviation values. In this test, we used a Hewlett-Packard 8640B signal generator and a SINADder. We measured the sensitivity of the receiver at the 12 dB SINAD point at deviations of 500 Hz, and 1 kHz through 10 kHz deviation in steps of 1 kHz. The results of the test are shown in **fig. 4**.

Notice that maximum sensitivity (−122 dBm at 12 dB SINAD) occurs at 3, 4, and 5 kHz deviation. The sensitivity is not affected by changes in deviation within this range. But above 5 kHz and below 3 kHz deviation, the sensitivity actually decreases. This result contradicts the popular notion that increasing the deviation of a transmitter increases range, and further indicates that reduction of transmitter deviation below 5 kHz does not reduce range (down to no less than 3 kHz, that is).

receiver selectivity

Although the performance of a receiver in rejecting

table 1. Level of isolation from interference experienced on channels separated from 10 to 30 kHz from an adjacent FM source.

Kenwood TW-4000A. On channel signal: −115 dBm, ±3 kHz deviation, 1000 Hz modulation.

interference modulation	30 kHz (dB)	20 kHz (dB)	15 kHz (dB)	10 kHz (dB)
400 Hz (EIA)	86	80	45	0
800 Hz	86	80	40	0
1200 Hz	86	80	33	0
2000 Hz	86	68	25	0
Kenwood TR-7800. On channel signal: −114 dBm, ±3 kHz deviation, 1000 Hz modulation				
400 Hz (EIA)	87	82	65	2
800 Hz	87	83	57	0
1200 Hz	87	83	48	0
2000 Hz	87	79	37	0
Handheld 1. On channel signal: −115 dBm, ±3 kHz deviation, 1000 Hz modulation				
400 Hz (EIA)	69	52	35	0
800 Hz	69	52	34	0
1200 Hz	69	52	30	0
2000 Hz	69	52	25	0
Motorola Syntor-X, 460.425 MHz. On channel signal: −107 dBm, ±3 kHz deviation, 1000 Hz modulation.				
400 Hz (EIA)	93	85	53	13
800 Hz	93	85	53	13
1200 Hz	93	85	50	20
2000 Hz	93	84	43	8

off-channel signals is something that cannot be adjusted easily, it is a major element of any radio communications system. A receiver consists of RF, IF, discriminator, and audio sections with most of the selectivity provided by the IF filter section. Intermodulation products and images can be generated in the RF and mixer stages. However, these are not directly related to the problem of adjacent channel interference — the IF filter and discriminator are.

Most FM receivers use crystal or ceramic filters to narrow the IF bandwidth before the signals reach the discriminator, where they are demodulated (back) to audio frequencies. While it would be nice if we could build ideal filters that would pass all signals in the desired passband and completely stop all off-channel signals, this isn't possible. Filters actually have finite passbands with "skirts" that roll off signals more the further away from the channel center frequency they are. The filters are usually specified by their bandwidth at the −6 dB and the −60 dB points; this is also how most ham transceivers are specified for selectivity.

Because the actual performance of the radio depends on this and other, less easily described factors — including discriminator performance — commercial manufacturers have therefore elected to specify their receiver selectivity with a functional test that actually challenges the receiver with a signal in the ad-

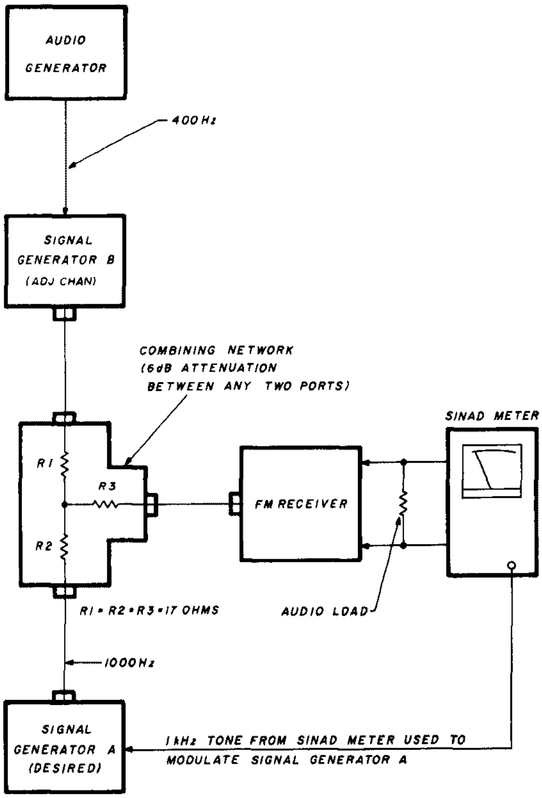


fig. 5. Setup is used to perform the adjacent channel selectivity tests. (From *Standard Radio Communications Manual* by Harold Kinley, © 1985, Prentice-Hall, Englewood Cliffs, N.J. 07632. Reprinted by permission of the publisher.)

jacent channel and measures the result. This is the test we selected and performed to determine selectivity.

The Electronic Industries Association (EIA) has established an adjacent channel rejection test based on the ratio between the on-channel to off-channel signal strengths when the received signal-to-noise and distortion (SINAD) ratio becomes degraded by 3 dB by the adjacent channel signal. This test, part of the RS 204-C test, is performed by mixing the signals from two signal generators and measuring the SINAD of a 1000 Hz tone modulating the on-channel signal at 3 kHz deviation.[4]

The test setup used to perform the selectivity test is shown in **fig. 5**. The on-channel signal level is raised to obtain a 12 dB SINAD, then raised an additional 3 dB. The off-channel signal is modulated at 3 kHz deviation by a 400 Hz tone, and its signal level is raised until the SINAD is degraded back down to 12 dB. Then the ratio of the two signals' strength is calculated in dB. When this measurement is made for both the next higher and the next lower adjacent channels, the lower of the two figures is used.

When the EIA established these tests for selectivity, they also established standards they consider "mini-

the action is at the IF — not the RF — stages

When the problem of adjacent channel interference is examined, attention is focused on the filtering that takes place at the intermediate frequency (IF) stages of the receiver, not at the radio frequency (RF) stages. The reason the IF gets the attention is the very narrow bandwidth required to allow separation of channels within the receiver's radio frequency input bandwidth.

At the RF frequencies, cavity resonators are usually used by repeaters and helical resonators are found in commercial and some consumer receivers. These filters are used to control the receiver's RF bandwidth to improve performance in terms of sensitivity and reduction of out-of-band signal strength. By this filtering, desensitization ("desense") and intermodulation distortion ("intermod") are reduced. However, these filters are typically 50 kHz to several Megahertz wide, and match the input RF stages to the intended operating range of the receiver. These filters are therefore very wide compared to the spacing of the channels (15 or 20 kHz), and will not have any significant filtering effect on those adjacent channels signals.

In the IF amplifier chain, however, the very narrow filters required become practical, due to both the lower frequency used in the IF (typically from 0.455 to 10.7 Megahertz) and the fact that the intermediate frequency does not have to be varied as the radio changes operating frequencies. In the IF stages, crystal filters are most commonly used to obtain very high "Q" (resonant frequency divided by bandwidth), frequency stability and shape factor (bandwidth at -60 dB divided by -6 dB bandwidth). These filters are commonly built with very narrow passbands (12 to 20 kHz wide for FM, and as little as 250 Hertz wide for CW applications). Even these filters do not act as "brick-walls," passing all signals in the passband and completely stopping all signals outside of the passband, since their out-of-band attenuation increases as the off-channel signal moves farther away from the passband. The slope of this attenuation is another factor in the response of a receiver to the adjacent channel rejection test, and together with the filter bandwidth (3 dB bandwidth) is a major factor in determining receiver performance in the test.

The IF filter, then, plays a key part in determining the receiver's response to adjacent channel interference, while the filtering at the RF stages of the receiver has little or no effect on this problem.

mum acceptable" performance. For this test, performed on the adjacent channels, the minimum acceptable standard is 70 dB isolation from the adjacent channel.

In these tests, we used a pair of HP 8640B VHF generators, chosen for their spectrally pure output signals (SSB phase noise below -130 dBc), as the signal sources. The SINAD was measured using a Helper Instruments "SINADder 5."

After the normal RS-204-C tests, we also measured selectivity with different frequencies of modulating tone on the adjacent channel signal. We did this because we believed that the choice of a 3 kHz deviation and a 400 Hz modulation tone may not be realistic for direct comparison with the ham environment, since our DTMF tones and voices contain higher frequency components than 400 Hz, and our transmitters may be adjusted for greater deviation. While we did not change the deviation setting, we made additional measurements with tones of 800, 1200, and 2000 Hz at 3-kHz deviation.

We measured receiver performance in this way, at channel spacings of 10 kHz, 15 kHz, 20 kHz, and 30 kHz. The seven units we tested included one commercial and three consumer mobile transceivers as well as three handhelds.

results with consumer gear

The results with consumer equipment are shown in **figs. 6A, B, and C.** Note that at 10 kHz spacing, little or no adjacent channel rejection is evident, and signals within 10 kHz of the channel center frequency are treated as "on-channel" by the receivers. This gives some idea of the bandwidth of each receiver's IF filter.

At 15 kHz separation, the adjacent channel isolation (of an unmodulated carrier) is about 45 to 70 dB. With the introduction of modulation, the interfering signal component is up by as much as 30 dB from ideal.

At 20 kHz, the adjacent channel isolation is about 80 dB, and some adjacent channel modulation is still detected. In most cases, the 20 kHz measurement was within a few dB of the receiver's ultimate rejection (as measured at 30 kHz separation).

At 30 kHz, the adjacent channel isolation is about 85 dB, and there is no change in this figure because of modulation frequency change. This figure shows little variation among the mobile rigs, but the handheld unit shows slightly lower performance (70 dB).(See appendix for further details.)

results with commercial gear

Motorola loaned us a commercial UHF "SYNTOR-X" which tuned to 460 MHz. (A VHF unit was not available.) At UHF, commercial manufacturers and Amateurs use 25-kHz channel spacing, but shop personnel believe that both VHF and UHF radios have similar specs and IF designs. *We believe this test is*

therefore representative of commercial receiver performance at VHF.

At 10 kHz, the SYNTOR showed slight rejection (see fig. 7), about 10 dB, of the interfering signal, indicating a slightly narrower IF filter than found in the consumer gear. Still, the low value means that receiver bandwidth is approximately 15-20 kHz total.

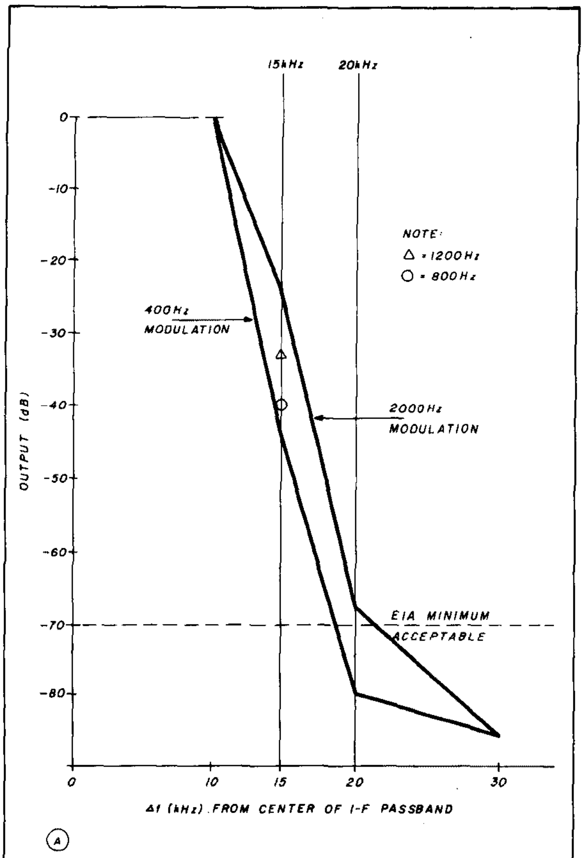
At 15 kHz, the SYNTOR showed 53 dB isolation, which was degraded by 10 dB when the modulating tone was increased to 2000 Hz. This again indicates, as in the case of the consumer gear, that we are on the skirts of the IF filter.

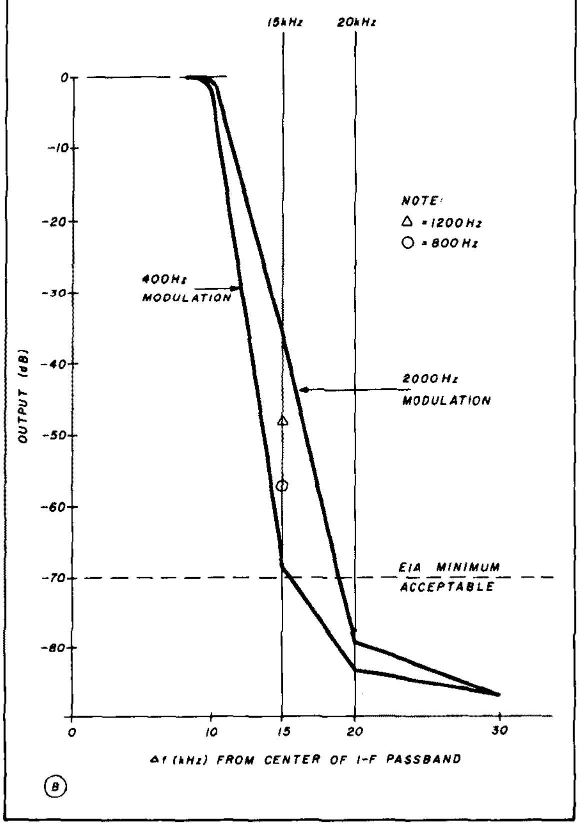
At 20 kHz, the isolation increased to 85 dB and was degraded only 1 dB by increasing the modulating tone to 2000 Hz.

At 30 kHz the SYNTOR showed 93 dB isolation, actually better than its specifications by several dB. Varying the modulating tone made no difference during the measurements.

discussion

Two major results are evident in this data. First, while commercial radio gear offers higher performance





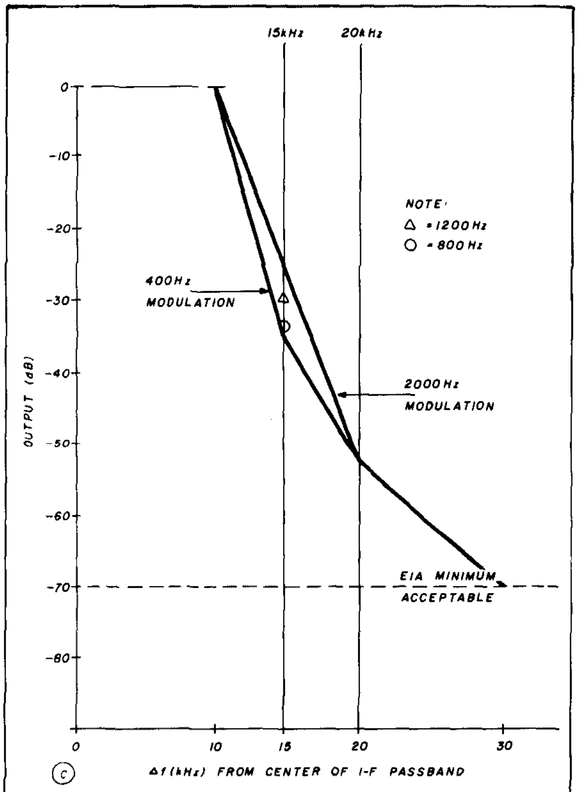


fig. 6. Effect of modulating frequency on selectivity test. Performed on three consumer (Amateur band) transceivers using 3 kHz deviation and 400 Hz (RS-204-C test), 800 Hz, 1200, and 2000 Hz modulation tones. *(A)* Kenwood TW-4000A, *(B)* Kenwood TR-7800, and *(C)* Handheld, HT1.

why do we use FM?

Considering the ongoing discussions of channel spacing and FM bandwidth, one might ask why hams use FM, which occupies such a large bandwidth compared with AM or single sideband (SSB). The answer lies in the improved signal-to-noise ratio (S/N) gained by the demodulator in an FM receiver. If you compare the signal-to-noise ratio of the demodulated signal with the carrier to noise ratio (C/N) of the radio wave before demodulation, you find that above a certain threshold, the demodulated signal shows a significant enhancement in S/N. The measurement of C/N must be made in a bandwidth equivalent to the IF bandwidth of the receiver, but within these constraints, we find an enhancement factor of: $E = 6\beta^2(\beta + 1)$ where β is the modulation index of the FM signal.*

To see how significant this enhancement is, consider the case of a 1000 Hz tone modulating a carrier at 4.5 kHz deviation, and a beta of 4.5, not unusual in Amateur voice systems. In this case, the enhancement is 668 times the carrier to noise ratio, or about 28 dB.

This enhancement, seen only above a threshold C/N, is one reason FM is popular for both commercial broadcasting and communications. Below this threshold, FM actually provides lower S/N than other modes, which is why weak-signal work is seldom done using FM.

*Simon Haykin, *Communications Systems*, John Wiley & Sons, 1983.

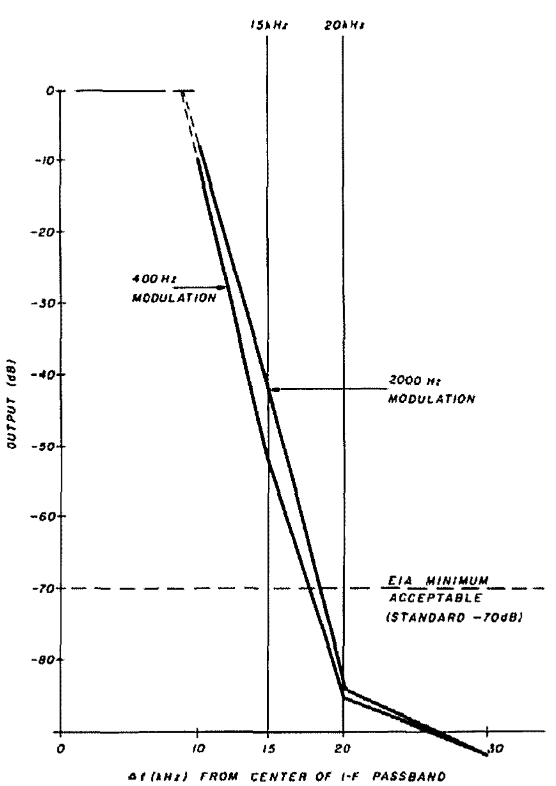


fig. 7. Selectivity test performed on a commercial UHF transceiver (Motorola "SYNTOR-X"). Both the EIA RS-204-C and 2000 Hz tones were used.

than consumer radios, the differences are not particularly large. Secondly, when operated at 15 kHz spacing, all these receivers will exhibit considerably degraded performance when compared to their use at 20 kHz.

On the first result, we wish to note that in the last several years, commercial radio suppliers have changed their radio designs from a relatively limited coverage radio to one that can cover channels separated by many Megahertz. This has been done by reducing filtering at the RF stages and enhancing the IF filters to maintain performance. While this reduction of the *Q* of the RF portion of the receivers does not alter the adjacent channel rejection, the enhancement of the IF stages does. The SYNTOR-X model is capable of covering the entire 450 MHz commercial band without retuning the RF stages, and represents first-class commercial radio equipment, with a price near $2800.

We expected, and found, excellent IF performance in the Motorola gear. The surprise was that the IF performance of the consumer gear *was actually quite similar*, and for most Amateurs the difference would not be significant — this was a surprise to us because we suspected that by adopting commercial standards for Amateur purposes, the interference problem could be solved. But the answer is clearly not that simple.

On the second result, we believe that when these radios are operated with 20 kHz channel spacing, they demonstrate performance which is near their ultimate design goal (as defined by their 30 kHz performance). At 15 kHz spacing, these radios *all* demonstrated very similar degradations in performance, and these degradations amounted to 30 to 40 dB. Furthermore, this degradation was significantly affected by the *bandwidth* of the interfering signal. Considering the conservative settings (3 kHz deviation, 400 Hz modulation) we believe the 15 kHz isolation numbers are generous compared to the Amateur environment, where 4.5 to 6 kHz deviation seems more common.

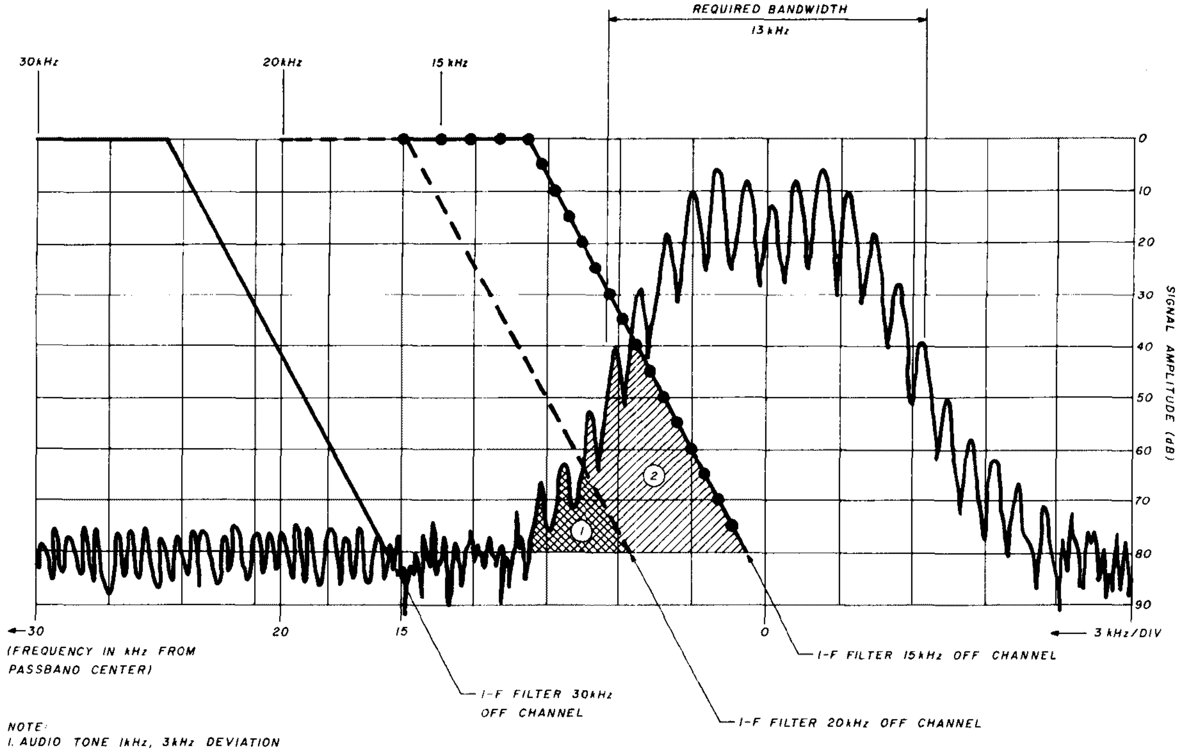


fig. 8. This graph clearly illustrates the need for at least 20 kHz separation between repeater channels.

Finally, when these results are compared with the EIA specification for minimum acceptable adjacent channel rejection, we see that all the receivers failed the test at 15 kHz spacing, and all but the handheld unit passed the test at 20 kHz spacing (see appendix).

The mechanism for this adjacent channel interference depends on both the nature of FM itself and the design of the receiver IF filters. What we believe is happening is shown in **fig. 8.** In this diagram, we have illustrated the shape of the FM signal resulting from a 1 kHz tone modulating a transmitter at 3 kHz deviation. First note the zone called "required bandwidth," which is the legendary 13 kHz wide. This zone shows the sidebands down to −40 dB from the carrier's unmodulated level. It is evident that some remaining sidebands are present, down to the −80 dB level, with the noise floor of the test instrument, an HP8568B spectrum analyzer.

On the left of the diagram, we have illustrated the filter shapes of typical consumer receivers spaced 30, 20, and 15 kHz away from the carrier frequency of the signal. Notice that at 30 kHz spacing, no power from the signal is entering the receiver's passband, down to the resolution of the instrument. At 20 kHz spacing, the edge of the receiver passband intersects a small portion of the signal, indicated by the area labeled *1*. At 15 kHz spacing, more of the signal is in the receiver passband, as noted by areas *1* and *2*. While it would be difficult to quantify the difference from this diagram, our tests have shown that this difference is in the range of 30 to 45 dB. If the more liberal EIA RS-204-C test were performed, using a 400 Hz tone, the receivers would pass at 20 kHz separation and *fail* at 15 kHz spacing.

We hope this report is informative and will be useful as you make your decisions on coordinating repeaters, both in frequency and geographical separation.

references

1. *Reference Data for Radio Engineers*, Sixth Edition, ITT and Howard W. Sams, Indianapolis, Indiana 46268, 1981, pages 23-7 to 23-9.
2. Carleton Maylott, W2YE, "Close Look at Amateur FM," *ham radio*, August, 1979, page 46.
3. R. Harold Kinley, *Standard Radio Communications Manual*, Prentice Hall, 1985.
4. *EIA Standard RS-204-C*, Electronic Industries Association, Washington, D.C., page 16, paragraph 14.2

appendix

Of the three handhelds tested (HT1, HT2, and HT3 — see table A1) variation existed from one unit to the other and also from the better side to the worse side of the filter. Because this test is made on the side of the filter (response) and not farther away from the passband, *the results will be very sensitive to the frequency to which the receiver is tuned*. Therefore, if the receiver drifts, it will change the better and worse isolation figures even more. There is also something else at work here: notice that HT3 shows considerable variation at 15 kHz from one side of the filter to the other, but its figures are more nearly equal at 20 and 30 kHz than HT1's. This is a dramatic demonstration of why a performance test like this is so much more revealing than merely quoting the nominal specification of the filter element.

table A1. RS-204-C test results for three handhelds (interfering signal, 400 Hz).

Channel Spacing:		15 kHz	20 kHz	30 kHz
dB isolation	(HT1)	32/50	48/70	69/77
	(HT2)	58/69	76/80	83/83
	(HT3)	23/66	70/73	76/77

This table is not meant to compare one brand name against the other (thus the anonymity) since none of the units are at their brand-new performance levels, but have been in use for varying lengths of time. Variation between individual units of a given model may also be considerable. To compare given models fairly, we would have to test several of each type to obtain a sample large enough to be considered representative of its series.

More important, though, is the difference in readings between the upper and lower adjacent channel tests and its effect on performance. Remember, the EIA (specification) procedure calls for using the lower of the two figures.

Finally, perhaps the EIA-70 dB specification is not meant for HTs; considering their intended use, it may be acceptable for HTs to have a lower isolation value since they are typically operated closer to the repeater and use a much lower gain and altitude antenna than found on most home or mobile installations.

ham radio

calculating the input impedance of a tapered vertical

Schelkunoff procedure — best method to use

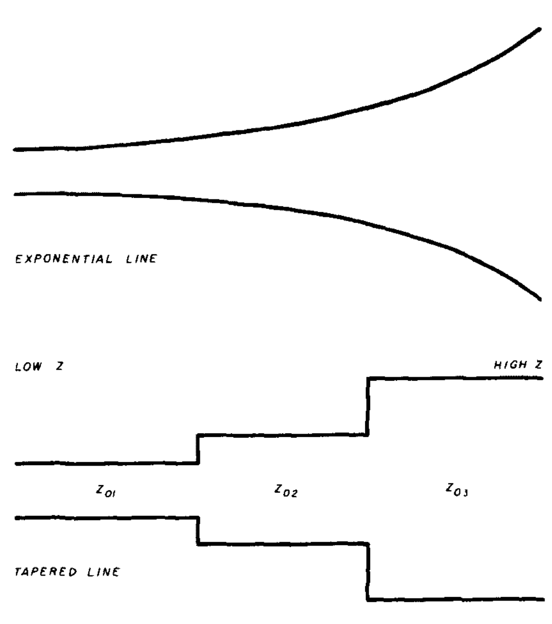


fig. 1. Comparison between tapered and exponential transmission line.

In the last decade there has been a tendency among Radio Amateurs to use vertical whip antennas on 80 and 160 meters. One clear advantage of using tapered whip elements, constructed of aluminum tubing, is their durability under severe weather conditions. But the fundamental question and the main obstacle to even more widespread Amateur use of the antenna is the problem of accurately determining the antenna feedpoint impedance for various tapers.

For cylindrical element length-to-diameter ratios, the feedpoint impedances are readily solvable by analytical methods such as Hallen's integral equation,[1] the induced EMF method,[2] and Schelkunoff's input impedance equation.[3] Using a computer and the "method of moments" approach, employing matrix algebra techniques,[4] will also provide answers.

which approach to use?

The basic question to be answered is what method should be employed to solve for the feedpoint impedance of a tapered vertical before its actual construction begins. Perhaps a clue is provided in the analytical method described by James Lawson, W2PV, in his well-known *ham radio* series on Yagi-Uda antennas.[5] Lawson stated that the inductance to capacitance (L/C) ratio of a tapered element is related to a geometrical mean diameter. Each cylindrical section has its own L/C ratio, or more specifically, its own surge impedance $Z_0 = \sqrt{L/C}$. This formula is used to describe an exponential (Collins) transmission

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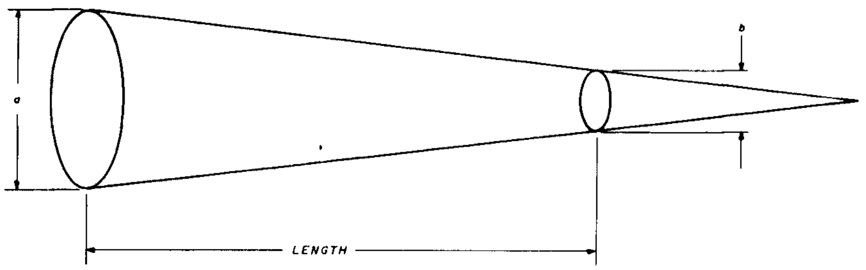

fig. 2. A tapered antenna element can be considered as a truncated cone.

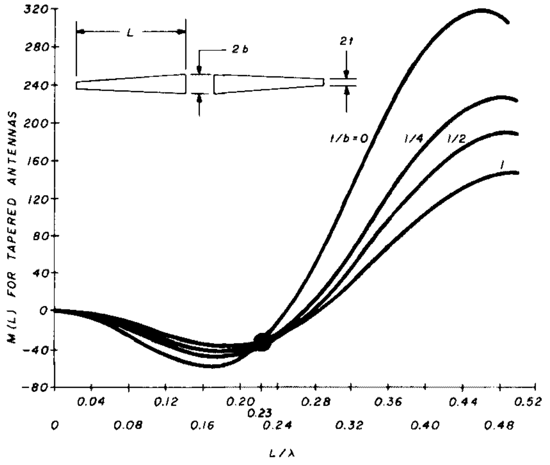

fig. 3A. "M" factor versus normalized length for tapered antenna elements. (Adapted from Sergei A. Schelkunoff's *Advanced Antenna Theory*, John Wiley and Sons, New York, 1952.)

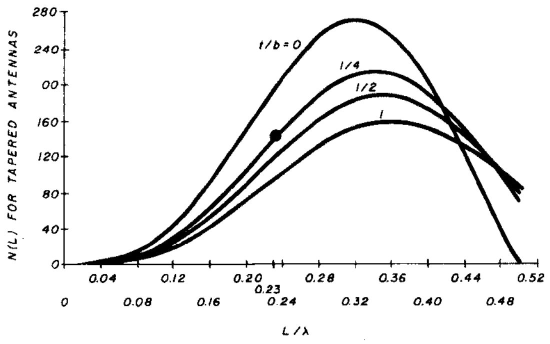

fig. 3B. "N" factor versus normalized length for tapered antenna elements. (Adapted from Sergei A. Schelkunoff's *Advanced Antenna Theory*, John Wiley and Sons, New York, 1952.)

line used for matching two different impedances (see **fig. 1**). Notice that each section of the exponential transmission line has its own surge or characteristic impedance. An impedance transformation occurs in either direction as a result of the taper. Consequently a tapered antenna element can be considered as a tapered exponential transmission line that has an average characteristic impedance described as a geometrical mean of its diameter.

W2PV took note of this physical relationship, applied that relationship to Yagi elements, and developed a method to evaluate the taper. An even earlier method, developed by Sergi A. Schelkunoff at Bell Labs is easier to work out — with either a pencil and paper or with a handheld calculator — yet still provides accurate results.

describing a tapered vertical

A tapered vertical element may be thought of as a truncated cone, as shown in **fig. 2**, with specific base and tip diameter and length, using cylindrical elements that telescope into one another, exhibiting impedance discontinuities at each section boundary. Consequently, **eq. 1** can be used to describe a tapered vertical element shape factor.[7]

$$Z_0 = 60\ \ell n(2L/b) + 60\ t/b - t\ \ell n(t/b) \quad \textbf{(1)}$$

where L = length (inches)
b = base radius (inches)
t = tip radius (inches)

Not only was Schelkunoff able to determine the shape factor, but he was also successful in making the necessary engineering approximations to equate antennas to transmission line behavior. This resulted in the modification of the basic transmission line formula for the solution of feedpoint impedance values for arbitrarily shaped antennas.[3]

$$Z_{in} = Z_0 \frac{R_a \sin G + j(X_a - N)\sin G - j(2Z_0 - M)\cos G}{(2Z_0 + M)\sin G + (X_a + N)\cos G - j R_a \cos G} \quad \textbf{(2)}$$

$$R_a = 60(\gamma + \ell n 2G - Ci 2G) + 30(\gamma + \ell n\, G - 2\,Ci\, 2G + Ci\, 4G)\cos 2G + 30(Si\, 4G - 2\, Si\, 2G)\sin 2G \quad \textbf{(3)}$$

$$X_a = 60\, Si\, 2G + 30(Ci\, 4G - \ell n\, G - \gamma)\sin 2G - 30\, Si\, 4G \cos 2G \quad \textbf{(4)}$$

where M and N values are taken from **figs. 3A** and **3B**

G = antenna height in radians

γ = 0.5772

Si = sine integral

Ci = cosine integral

example

It is desired that a vertical whip antenna is to operate on 3.8 MHz. Its height is 60 feet (18.29 meters) or 720 inches (1829 cm). The base diameter is 3 inches (7.62 cm) and its tip diameter measures 3/4 inch (2 cm). What is the vertical feedpoint impedance at this operating frequency? It is assumed that the vertical sits on a perfect ground radial system that has a minimum resistance.

Step 1. Find the shape factor (Z_0).

$$Z_0 = 60\, \ell n \quad 2 \cdot \frac{720}{1.5} \quad + 60 \quad \frac{0.375}{1.5} \quad - 0.375 \cdot \ell n \quad \frac{0.375}{1.5} = 384.3$$

Step 2. Find the M and N values from **figs. 3A** and **3B**.

$$\text{with } t/b = \frac{0.375}{1.5} = 0.25$$

$$M(1) = -38 \text{ and } N(1) = +125$$

Step 3. Express fractional wavelength of antenna in degrees or radians.

$$\textit{fractional wavelength} = \frac{height}{984/Freq\ (MHz)}$$

$$= \frac{60}{(984/3.8)} = 0.2317$$

$$0.2317 \cdot 360° = 83.4°$$

$$\text{or } \frac{83.4°}{57.3°} = 1.4557 \text{ radians}$$

Step 4. Calculate R_a value.

$$R_a = 60[0.5772 + \ell n\, 2.9114 - (+0.15)]$$
$$+30[0.5772 + \ell n\, 1.4557 - 2(+0.15) + (-0.10)]\cos 2.9114$$
$$+30(1.4356 - 2 \cdot 1.8431)\sin 2.9114 = 58.2$$

Step 5. Calculate X_a value.

$$X_a = 60(1.8431) + 30(-0.10 - \ell n\, 1.4557 - 0.5772)\sin 2.9114$$
$$-30 \cdot 1.4356 \cdot \cos 2.9114 = 145.31$$

Step 6. Calculate the vertical feedpoint impedance.

$$Z_{in} = 384.3 \left[\frac{58.2 \sin 1.4557 + j(145.3 - 125)\sin 1.4557 - j(2 \cdot 384.3 + 38)\cos 1.4557}{(2 \cdot 384.3 - 38)\sin 1.4557 + (145.3 + 125)\cos 1.4557 - j58.2 \cos 1.4557}\right]$$

$$Z_{in} = (384.3 + j0) \left[\frac{57.8 - j72.5}{756.8 - j6.7}\right] = (384.3 + j0)(0.0772 - j0.0951) = 29.7 - j36.5$$

$$= 29.7 \text{ ohms} - j36.5 \text{ ohms}$$

Using an HP-41C calculator eliminated the manual labor and produced answers more quickly (see **Appendix**).

This program can be used not only on tapered vertical antennas, but on tapered Yagi elements as well. To do this, calculate half the dipole length just as if it were a vertical, and then multiply the answer by two. This new product is the feedpoint impedance for a tapered dipole element.

```
01♦LBL "ZB"
02 XEQ "Ra"
03 RCL 05
04 SIN
05 *
06 STO 16
07 CLX
08 XEQ "Xa"
09 RCL 08
10 -
11 RCL 05
12 SIN
13 *
14 ENTER↑
15 XEQ "O"
16 RCL 07
17 -
18 RCL 05
19 COS
20 *
21 -
22 STO 17
23 CLX
24 XEQ "Ra"
25 RCL 05
26 COS
27 *
28 -1
29 *
30 STO 14
31 CLX
32 XEQ "O"
33 RCL 07
34 +
35 RCL 05
36 SIN
37 *
38 ENTER↑
39 XEQ "Xa"
40 RCL 08
41 +
42 RCL 05
43 COS
44 *
45 +
46 STO 15
47 CLX
48 RCL 17
49 RCL 16
50 RCL 14
51 RCL 15
52 XROM "C/"
53 RCL X
54 STO 18
55 CLX
56 RCL Z
57 STO 19
58 CLX
59 0
60 ENTER↑
61 XEQ "ZO"
62 RCL 19
63 RCL 18
64 XROM "C*"
65 "ZB IN=R"
66 AVIEW
67 RCL X
68 STOP
69 CLX
70 "J="
71 AVIEW
72 RCL Z
73 STOP
74 END
```

```
01♦LBL "ZO"
02 RCL 09
03 RCL 10
04 -
05 1/X
06 RCL 10
07 *
08 60
09 *
10 ENTER↑
11 RCL 10
12 RCL 09
13 /
14 LN
15 *
16 ENTER↑
17 RCL 11
18 2
19 *
20 RCL 09
21 /
22 LN
23 60
24 *
25 +
26 STOP
27 END
```

```
01♦LBL "O"
02 XEQ "ZO"
03 2
04 *
05 RTN
06 END
```

```
01♦LBL "Ra"
02 RCL 04
03 LN
04 .5772
05 +
06 RCL 06
07 CHS
08 +
09 60
10 *
11 ENTER↑
12 RCL 05
13 LN
14 .5772
15 +
16 RCL 06
17 2
18 *
19 CHS
20 +
21 RCL 03
22 +
23 RCL 04
24 COS
25 *
26 30
27 *
28 +
29 ENTER↑
30 RCL 01
31 2
32 *
33 CHS
34 RCL 02
35 +
36 RCL 04
37 SIN
38 *
39 30
40 *
41 +
42 STOP
43 RTN
44 END
```

```
01♦LBL "Xa"
02 RCL 03
03 RCL 05
04 LN
05 -
06 .5772
07 -
08 30
09 *
10 RCL 04
11 SIN
12 *
13 ENTER↑
14 RCL 01
15 60
16 *
17 +
18 ENTER↑
19 RCL 04
20 COS
21 RCL 02
22 *
23 30
24 *
25 CHS
26 +
27 STOP
28 RTN
29 END
```

fig. A1. HP41C calculator program for finding feedpoint impedance of tapered vertical antenna.

appendix

HP-41C calculator program for finding feedpoint impedance of tapered vertical antenna

HP-41C instructions

- Execute SIZE and key in 020.
- Key in ZB program and subroutines Z_0, O, R_a, X_a.
- Key in functions memory register 01 to 11.
- Execute program ZB.
- Each subroutine answer will be displayed, to continue program depress R/S key.

functions	memory registers
Si 2G	STO 01
Si 4G	STO 02
Ci 4G	STO 03
2G	STO 04
G	STO 05
Ci 2G	STO 06
M function	STO 07
N function	STO 08
base radius	STO 09
tip radius	STO 10
length	STO 11

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5. James L. Lawson, W2PV, *ham radio*, January, February, May, June, July, September, October, November, December, 1980. (*For limited time only a set of nine back issues containing W2PV's series of articles on yagi antenna design will be made available at the reduced price of $9.86 postpaid to U.S. addresses, $16.95 to addresses outside the U.S. Foreign payment accepted in U.S. funds drawn on U.S. bank only. Address request to Ham Radio's Bookstore, Greenville, NH 03048. Supplies are limited; order promptly.*)
6. James L. Lawson, W2PV, "Yagi Antennas: Practical Designs," *ham radio*, December, 1980, page 33.
7. Sergei A. Schelkunoff and Harald T. Friis, *Antennas: Theory and Practice*, John Wiley and Sons, New York, 1952, pages 425-431.

ham radio

August 1985

short circuit

In "Voltage Controlled Oscillator Uses Ceramic Resonators," (K2BLA, June 1985, page 23), two base bias resistors for the 2N3904 were inadvertently omitted from fig. 8. Add a 2.2k resistor from base to ground and a 10k resistor from the same base to +15 VDC.

design a
toroidal tank circuit
for your vacuum tube amplifier

No engineering degree required for this practical approach

Sooner or later every ham experiences three common desires: to build a big low-band power amplifier covering all of the ham bands, including 160 meters; to acquire (without benefit of an EE degree) sufficient knowledge to determine the exact number of turns and placement of taps for the tank coil therein; and to see, sometime soon — preferably within the limits of this century — the shrinking of the tank circuit to an acceptably small size.

Good news! With no more than a high school student's understanding of math, you can easily fulfill all three desires described. How? By simply using a powdered iron toroidal core network in the vacuum tube tank circuit in place of the huge conventional air dielectric coil.

Many solid state amplifiers, by virtue of their extremely low load impedances, use this technique; this is one reason for their reasonably small size. Yet only a few experimenters have actually tried using the toroid in the high impedance circuits of the vacuum tube.

After reading this article, you'll be able to either modernize the amplifier you have on the shelf or start from scratch, building one from the many excellent schematics available in periodicals, in Bill Orr's *Radio Handbook,** or in the various ARRL publications.

What effect does continuous high power have on the core? The high impedance of vacuum tubes causes the toroid core to become hot, causing its characteristics to change. The core would also saturate at the higher frequencies, causing instability and destruction of its composition. But the core described in this article will easily handle in excess of 2 kW maximum peak power, with no instability or saturation. (When I finished testing, the core was barely warm to the touch. Before you touch the coil, be sure the B + is off and capacitors fully discharged to avoid shock.)

The toroid used was an Amidon T-400-2A. A nearly exact substitute for this is two Amidon T-400-2 units sandwiched together. If the amplifier you want to build is less than the 2 kW class, then a single T-400-2 core can be used. Even more space can be saved by using a pair of 3-inch (7.6 cm) T-300-2 toroids. These, however, will require five or six more turns of wire.

choosing the right tube is an important first step

You must now decide on the type tube or tubes you intend using in your amplifier. Grounded grid operation in class AB1 is a favorite for single sideband. Let's suppose you have a number of 4CX300A tubes and sockets or 4CX250Bs or Fs that were obtained surplus. The very first number you must come up with is the operating plate load impedance. This you determine

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*Available from Ham Radio's Book Store, Greenville, New Hampshire 03048, $12.95 plus $3.50 shipping and handling.

from the level of plate voltage and current you'll be running. This information can be obtained from the sources previously mentioned or from data sheets from the tube manufacturer.

Suppose, for example, that SSB is your main interest and a pair of 4CX300A tubes is available. We've already stated that grounded grid would be the choice. But the 4CX series of tubes requires screen and grid bias voltages, so to call the operation "grounded grid" would be inaccurate. The tubes would really be running in a cathode driven circuit with the grids at RF ground potential, but above ground at DC so that the necessary voltages could be applied.

With drive applied in class AB operation, each tube draws 250 mA at a plate voltage of 2000 volts. This translates to a *total input power level of 1000 watts.* Output efficiency in this class of service is between 60 and 65 percent, or just over 650 watts. A circuit Q of 12 is needed to sustain the proper energy storage in the tank circuit.

determining plate load resistance

We now have all of the information necessary to calculate one of the more important values needed, and that is the load impedance, R_p, the tubes will present to the input of the Pi network. For class AB operation

$$R_p = \frac{V_p}{1.8 \cdot I} \quad (1)$$

where V_p = plate voltage
I = total plate current

$$\frac{2000}{1.8 \times 0.5} = \frac{2000}{0.9} = 2222 \text{ ohms}$$

For class "B" operation the formula becomes

$$R_p = \frac{V_p}{2\ I_p} \quad (2)$$

Values of components C1, C2, and L1 in a pi configuration are readily available.[1] An abbreviated version of it is provided in **table 1.** *The data is divided into* columns headed by the various values of plate load resistance in increments of 250 ohms. The tables are usually calculated for a Q of 10 or 12. C1 is called out as the plate tuning capacitor and C2 as the loading capacitor. The purpose of the pi networks is to step down the high level of plate load impedance to some value between 25 and 100 ohms to match the load, which is usually an antenna or possibly a dummy termination.

pi values for other load impedances

The plots provided in **figs. 1, 2,** and **3** can be used to find component values for intermediate load impe-

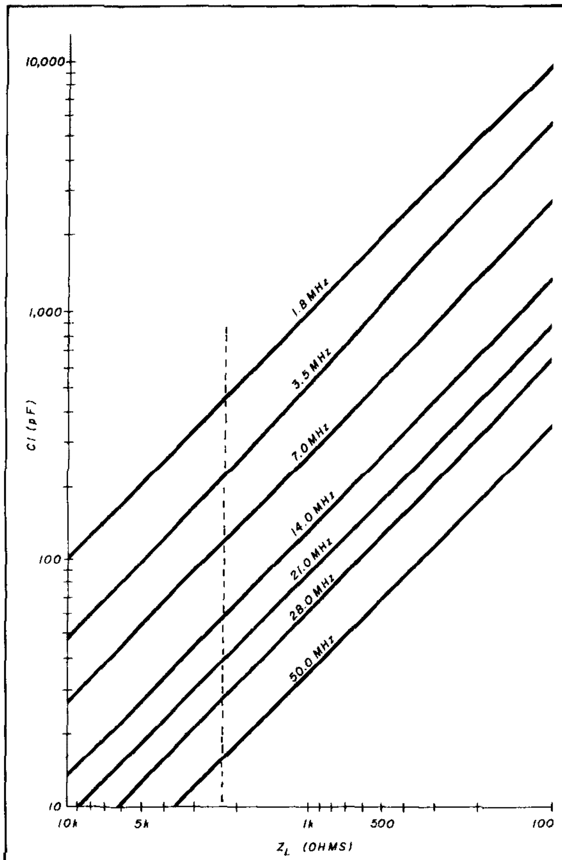

fig. 1. Plot of plate tuning capacitor, C1, plate load impedance (Q = 12).

dances simply by linearly interpolating. For example, if the value of capacitor C1 is needed for a load impedance half way between 1500 and 2000 *ohms on the* 160 meter band, take the arithmetic mean:

$$C1 = \frac{531 + 430}{2} = 481 \text{ pF}$$

The same holds true for determining the values of C2 and L1. The same interpolation method can be used to find pi values for the WARC band frequencies.

effective pi-network capacitance

The two capacitors, C1 and C2, in the pi-network, are effectively in series and shunt the coil so that the resultant capacity (C_T) of the pair determines the resonant frequency (see **fig. 4**). Knowing C_T and the required resonant frequency, the coil value can be calculated. Evaluating:

$$C_T = \frac{C1 \cdot C2}{C1 + C2} = \frac{481 \times 2652}{481 + 2652} = 407 \text{ pF}$$

remember this, because we'll be using it later.

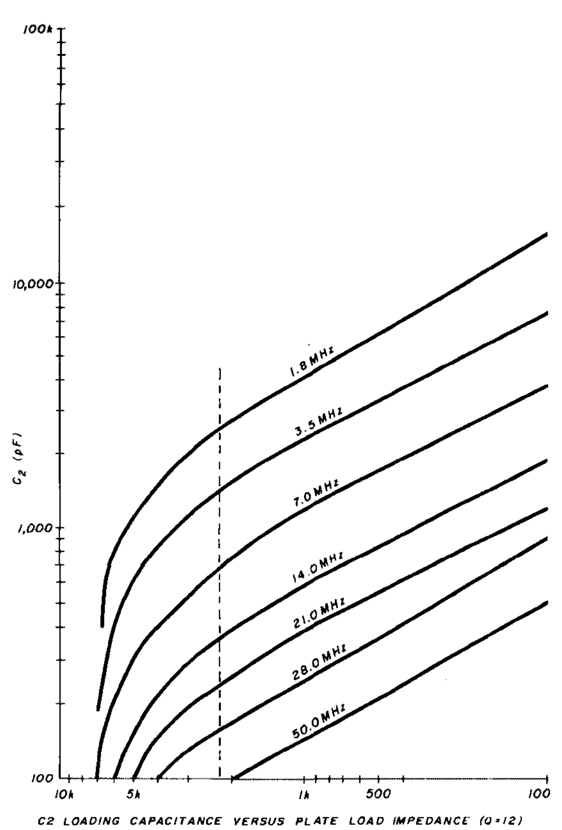

fig. 2. Loading capacitance versus plate load impedance (Q = 12).

table 1. Pi-network component values versus plate load impedance. (Capacitors C1, C2 in pF and inductance L1 in microhenries.)

	Z_L plate load impedance (ohms)			
capacitor C1				
band	**1750 ohms**	**2000**	**2250**	**2500**
160	610	531	481	430
80	318	273	246	220
40	159	136	123	110
30	120	102	93	83
20	80	68	62	55
15	53	45	41	37
10	40	34	32	30
capacitor C2				
160	3176	2865	2652	2440
80	1628	1473	1368	1263
40	815	737	684	632
30	610	492	457	422
20	407	368	342	316
15	272	246	229	211
10	204	184	171	158
inductor L1				
160	14.94	16.61	18.36	20.10
80	7.56	8.54	9.72	10.90
40	3.78	4.27	4.89	5.50
30	2.52	2.84	3.24	3.64
20	1.89	2.14	2.42	2.70
15	1.26	1.42	1.62	1.82
10	0.95	1.07	1.21	1.36

toroids in general

Information about toroidal cores and coils is available in many handbooks. Amidon's catalog, "Iron and Ferrite Cores," also includes general information about core characteristics.

Toroidal cores are basically of two types: powdered iron and ferrite. The permeability of the core (μ) helps determine the number of turns of wire required on a given physical size core; the larger the μ, the fewer the number of turns required to provide the given inductance. Powdered iron cores generally have permeabilities from 1 up to 125, while ferrite cores have permeabilities ranging from 40 to 5000. Worthwhile noting is that with toroids, there is no such thing as a partial turn. If the wire goes through the hole, you have one turn and you do not get a second turn until the wire goes through the window once more.

The core size is another factor that affects the number of turns needed for a given inductance. This is directly related to the cross sectional area and consequently the flux density of the core. It sounds complicated, but is simplified by combining everything into what is called the "A_L" value, and this alone can be plugged into a simple formula resulting in the number of turns required for a given inductance. Although the ferrite material has a much higher μ than the powdered iron core, ferrite is not as stable and saturates easily when used in power circuits.

Certain factors must be known in order to select the proper core for a specific job; first, we need the frequency range and permeability. Like resistors, all cores are marked according to a universal color code representing the compound mix number (**table 2**). The compound mix determines the core's frequency range. For the HF range, compound mix number 2 with a permeability of 10 is best. The core is powdered iron and is colored red.

Toroidal cores are prefixed either with a *T* for powdered iron or an *FT* for ferrite. The number following the *T* or *FT* identifies the diameter of the core. One of the more popular cores — because of its extensive application in antenna baluns — is the T-200-2, which is short for "toroidal powdered iron, 2 inches (5 cm) in diameter, No. 2 compound mix." **Table 2** lists core colors and mixes versus frequency.

selecting the core

The core I chose is a T-400-2A, which has an A_L

of 360. As stated previously, two T-400-2 are a nearly exact equivalent; each 400-2 has an A_L of 185. Two stacked, therefore, would have an A_L of 370, because two of these together are a little thicker than a single 400-2A. You've probably guessed that a suffix letter relates to a core with a somewhat thicker than standard width and consequently a higher A_L. (Why the A_L number? This number represents the inductance in microhenries that 100 turns will produce on any given core.) From this important factor one can determine the number of turns required for a specific inductance.

The A_L number is used in a simple formula:

$$Turns = 100\sqrt{\frac{desired\ inductance\ (\mu H)}{A_L\ value}} \quad (3)$$

The coil number of turns is calculated for the lowest band we are to use, which is 160 meters. Referring to the capacity (inductance data) we find, under 2250 ohms, that the inductance for 160 meters will be 18.36 — let's say 18 μH. Plugging in the numbers,

$$100\sqrt{\frac{18}{360}} = 100\sqrt{0.05} = 22\ turns$$

In case of a fraction, round out the number, since there are no partial turns in winding a toroid.

As a point of interest, I noticed that Amidon now has a 3.048 inch (7.8 cm) core called a T-300; A-2 mix would have an A_L of 115. The core is 0.5 inch (12.7 mm) thick. Stacking one on top of the other would make up a 1 inch thick core with an A_L of 230. The required number of turns in this case is equal to

$$100\sqrt{\frac{18}{230}} = 27.975\ or\ 28\ turns.$$

For those of you thinking of an amplifier in the 1 kW input class, the coil should be wound with No. 12-gauge enamel covered wire. If you contemplate 2 kW, use No. 10-gauge wire.

coil preparations

Obtain a roll of glass cloth electrical tape No. 27. It costs approximately $3.50 for 66 feet, or about 20 meters (there'll be plenty left over). If you're going to stack cores, use Eastman 910 super glue or equivalent. Align the cores, applying a small quantity of the glue — possibly only a few drops — around the side of one core. Work swiftly; super glue hardens quickly. The glue prevents the cores from moving out of alignment while they're being prepared for winding.

Now wrap two layers of glass tape around the core. Apply a heavy layer of Polystyrene Q-dope (available from Radio Shack for less than $2 per 2-ounce bottle) to the glass-taped core. Set the core on a sheet of waxed paper or plastic wrap. After 20 minutes or so the Q-dope will have hardened sufficiently for another layer to be applied. (The core can easily be separated from the waxed paper or plastic wrap for full coverage.) When you're finished, no one — not even *you* — will know if it's one core or two.

table 2. Color code identifies core mix and frequency range.

frequency range (MHz)	color	mix
0.05- 0.5	gray	3
0.1 - 1.5	red/white	15
0.5 - 5.0	blue	15
1.0 -30.0	red	2

Apply coat after coat until you've practically exhausted the contents of one 2-ounce bottle of Q-dope; you'll probably have about ten layers, which will add about 1/8 inch (.32 cm) or more to the thickness of the core. Now you'll need to add some insulation to the sides of the toroid. Some 1/16 to 1/8-inch (16 mm to 32 mm) polystyrene or fiberglass will be needed. (Don't use plexiglass.) In my search I ended up using 1/16-inch (16 mm) G-10 epoxy glass printed circuit board.

Use a propane torch to remove the copper foil. Place the board in a vise and apply the heat, stripping off the copper foil with a long-nose pliers. It takes only a few minutes to strip both sides when using a torch. Use of a fly cutter (circle cutter) to cut two identically-sized donut washers out of the epoxy board. The outside diameter should be 1/4-inch (6.4 mm) wider than the epoxied core, and the inside window or center hole 1/4-inch (6.4 mm) smaller.

Using some of the remaining coil dope, coat one side of each of the donuts you've prepared and place one on each side of the powdered iron core. Let it set. Now when you place the wire winding on the core, the wire will clear the core by 1/16 inch (1.6 mm) or 1/8 inch (3.2 mm) and there'll be plenty of insulation on the sides of the core. The Q-doped glass tape and end plates are the instruments that prevents RF from arcing to the core. If you prepare the core as directed, this will be no problem

If you're using a deep chassis the coil can be held in place by the wire leads and mounted directly to the switch, wheel fashion. My chassis was not deep enough, since it was subdivided with the input of the RF amplifier on one side and the output on the other. I prepared two more pieces of G-10 epoxy as large washers, cutting a small hole in the center of each so that the core could later be mounted from a long stud secured to the chassis. One recommended step you can take is not absolutely necessary, but does make

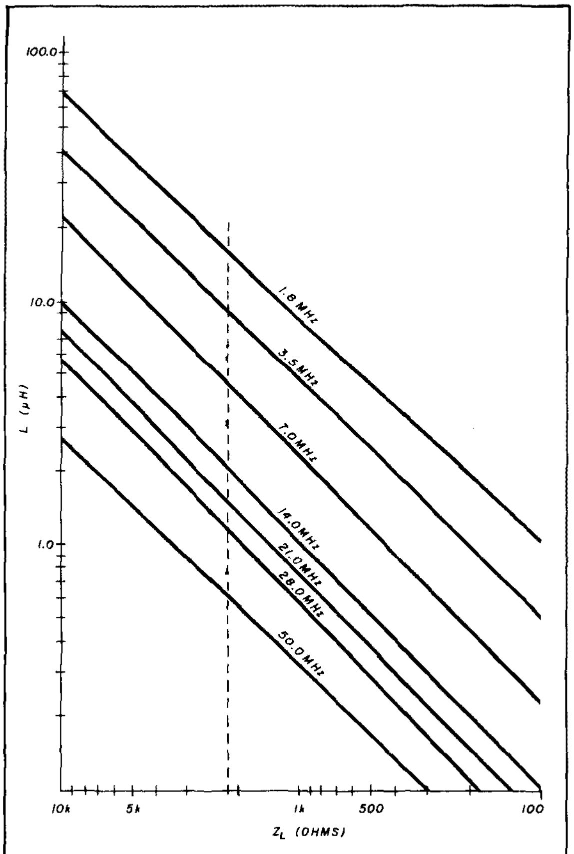

fig. 3. Plate inductance versus plate load impedance (Q = 12).

a neater package. The core does not include a 10-meter winding. As is conventional, this coil is almost always separate and is made up of about three or four turns of 1/4-inch (6.4 mm) copper tubing 2 to 2-3/4 inches (5 to 7 cm) in diameter. I mention it at this point because you may already have such a coil and the toroid core should be wound in the same direction as the turns of 10-meter coil, in order to prevent possible problems later. Take the core and determine the direction of winding, applying a sample of small wire to determine how the wire will lie in the core. With a Swiss file, file notches for the winding, making them about 3/64 inch (1.58 mm) deep. If you have 20 turns, there will be 40 notches on the inside window circumference and another 40 on the outside diameter of the G-10 epoxy donut sides.

Some pointers on winding the core: don't put the core between the jaws of a vise; it may break. Apply a test turn of wire to the core, observe the point at which a full turn is completed, remove and measure the test turn, multiply that figure by 20 and add 1 foot (30.48 cm). This is the length of wire you'll use to wind the core. Place one end of the wire in a vise, go to the far end and get a good grip on the end with large pliers, stretch the wire taut and tug sharply to remove any kinks. (I didn't have anyone to help me at this point, but I suggest you get someone to hold the core while you wind it.) Put the end of the wire through until the core is close up to the vise. (Ask your helper to hold it.) In placing the notches around the core for the wire, wind till no gap remains. Leave a space between the beginning and end of the winding equal to at least the space of 1 or 2 turns of wire. This also provides a starting point for the winding. If you start on the notches for the second turn you can get the last turn on at the beginning from the end of the wire that was held in the vise. Start threading the wire. The person holding the core can assist in holding the wire in place. When finished, tie the ends together temporarily so they will not unravel or become loose. Some coil dope can be used to secure the wire in the notches.

determining the tap positions

As the wire is spaced from the core, it's easy to scrape the enamel from the wire. Use the A_L formula to determine the tap placement. I'll take you through the first tap determination — that of 80 meters. Referring to the inductance data for L1, 80 meters = 9.72 μH.

$100\sqrt{\frac{9.72}{3600}} = 100\sqrt{0.027} = 16.43$ so the tap is 6.5 turns from the 160 meter end or 16.5 turns from the load end. It can be placed at 16 turns. Calculate the location of each tap for the remaining bands.

If you wish — and it might be a wise thing to do — prove you have made all the correct calculations. How? You'll need a grid dipper for this part. Most of this can be done prior to winding the coil in its finished form. You can wind a more manageable wire size of No. 14 or 16 to test for the correct value of inductance.

At the beginning of this article we determined that 407 pF was needed for C1 and C2. With 22 turns of No. 14 or 16 wire on the core, make the last turn loose so that you can couple a grid dipper to it. Using capacitors from your junkbox, make up a capacitor of approximately 350 pF (300 and 50 pF are standard values of dog bone dipped micas). Connect the capacitor across the two ends of the winding — making sure the coil is not placed on a metal bench — couple it to the last turn, and dip the meter. If the dip is broad, de-couple by backing the dipper away until you get a sharp dip. Read the dipper frequency. If the scale has poor resolution, couple it to a counter or a receiver.

Many dippers only go down to 2 MHz. Wrap the leads of a 10 or 20 pF capacitor around the pins of the dipper coil. This will extend the range of the dipper, but the readout will be null and void so listen for the dipper oscillator on your receiver reading the receiver dial or digital readout. If this is not possible, compute the resultant capacitance for 80 meters and just place the equivalent capacitance across the 14 turns, dip the meter, and read the dial. Then place a one-turn link through the coil and tie the ends of the one turn together. Couple the dipper to this one turn. It will get you close to the frequency. It won't be as accurate as coupling to a loosely wound turn of the coil itself because a shorted turn loads the coil somewhat.

tricks of the trade

Looking at the capacitor values of C1 and C2 for 160 and 80 meters makes one gasp. I used a four-gang capacitor from a surplus Hewlett Packard audio oscillator in my 800 watt output amplifier for C2. This has a total capacity of 2100 pF. I used two positions on the bandswitch for 160 and switched in 2000 pF on the lower frequency and 1000 on the higher end. The fixed capacitors used were the Hi-Q CRL-850 series 5000 volt DC.

For the plate capacitor C1, I also switch in parallel capacitors. Let's take a second look at the C1 values for 160, 80, and 40 meters 481, 246, and 123 pF respectively. We'd like to keep the physical size of the amplifier down, so let's say that a realistic practical maximum value is 250 pF. If one is to use vacuum variables, 350 becomes practical but for an air dielectric variable 250 pF seems high enough.

combining capacitors

How can we get by with using 250 pF when we need 481 pF for 160 meters and 246 for 80? How would one cover the band? We may need two positions on the bandswitch if we plan to tune the whole 160 meter band, but let's see what it takes. For an accurate frequency plot it would be nice to have a reactance slide rule, but we can come close enough by interpolating between 3.5 MHz and 1.8 MHz in **figs.** 1 and 2. We know the 160-meter inductance is 18.4 μH and the resultant capacitance (C_T) of C1 and C2 is 410 pF for the low frequency end (1.8 MHz). What capacitance is required for 2.0 MHz? We find that it will take 340 pF. This is a change in the resultant capacity (C_T) of 70 pF over the band. If C_T = 340 pF, what will be the capacity remaining in the plate tuning capacitor C1? Let's go back to the plot in **fig. 2** for C2 at 2250 ohms load impedance. Extending a line through this impedance point, we find that 2400 pF will be required at 2.0 MHz — the high end of the band. Let's now look at the chart for C1; it looks like 400 pF. Let's see when

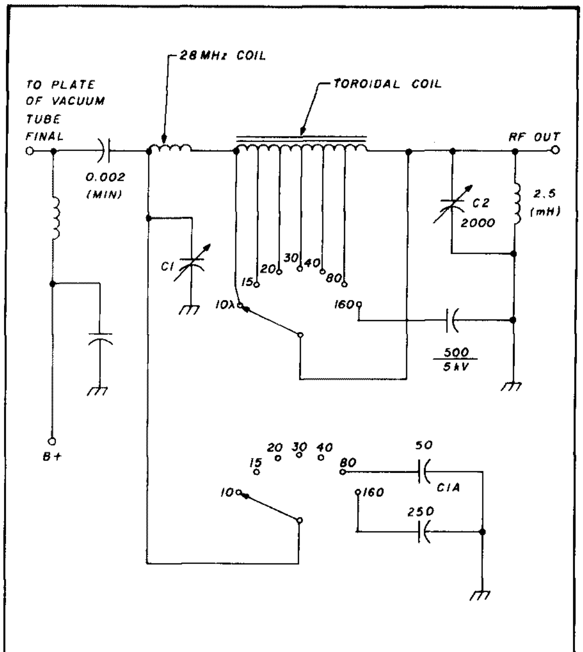

fig. 4. Final pi-network design works on all bands including WARC frequencies.

evaluating for the series capacitance how close we come to the required 340 pF. $\frac{2400 \times 400}{2400 + 400} = 342.9$ pF.

That's close enough.

This indicates that C1 must be variable from a high of 481 pF for 1.8 MHz to a low value of 411 pF for 2.0 MHz. From this, if we switch in 250 pF fixed capacity across our 250 pF C1 selected variable, there will result a 500 pF total maximum capacitance which is 19 pF more than needed. For the high frequency end, if we tune out, say, 70 pF of the 250, there will be 430 remaining in the circuit. We still have a 180 pF of variable capacity remaining. It looks as if we can do the entire 160-meter band with just one band position by switching in 250 pF of CRL-850 series capacitance.

Let's now look at 80 meters and evaluate it the same way. 246 pF is required for C1 and 1368 pF is necessary for C2. We have plenty of C2 with 2100 pF of variable, but we may need more C1 — about 50 pF more. First let's see the minimum we need to cover 4.0 MHz. The capacitance chart at 2250 ohms, although poor in resolution, looks like 230 or possibly 240 pF. Let's say we switch in 50 pF of the CRL-850 series capacitance. This will provide a variable capacitance of 300-250 + 25 (about 25 pF for minimum capacitance in the capacitor and circuit capacity). This equates to a minimum available capacitance of 75 pF and a maximum of 300. We need only 246, so we have an excess of 54 pF; we need

a minimum of 230 or 240, and we have 175 — so we're well in on 80 meters by switching in 50 pF. All the other bands are OK without additional capacitance. This should be enough data to enable you to draw up a switching circuit — see **fig. 4**.

bandswitch connections

The bandswitch should be of ceramic or porcelain material, of high quality and capable of handling the RF power. The switch will need two decks to accommodate the additional C1 capacitance required to be switched in and out. The number of positions will depend on the number of bands.

Some final thoughts: keep in mind that if the load impedance of the vacuum tube or tubes selected calculates out to be 1500 ohms or less, the required L1 is somewhat less, so C1 and C2 become larger. This means there will be more capacitance to switch in. With such high values of fixed capacitance there may not be enough variable capacitance left to cover the 160-meter band. If this is the case, two positions will have to be provided on the bandswitch in order to cover both the low and high frequency segments of the band.

Note that only seven positions are shown on the switch in **fig. 2** and the bands for 18 or 24.5 MHz are omitted. Also note that to provide enough Q on the 10-meter band, a separate coil will be required the same as in any and all amplifiers on the market. This usually requires about three or four turns of 2 to 2-1/2 inch (5 to 7 cm) diameter 1/4-inch (6.4 mm) copper tubing.

The last item is the plate blocking capacitor. This 0.002 μF capacitor is an important item and must handle high current, is high Q like the CRL-850 series capacitors and also is high voltage about 10,000 volts in amplifiers of 2 kW. This value capacitor has a 45-ohm reactance on 160 meters. The 0.004 μF capacitors reactance value is 25 ohms on 160 meters. The 858 capacitors are rated at 5000 volts and require two 1000 pF units in parallel. The 2000 pF capacitors are only rated at 1.5 kV. The 5000-volt 858 is just marginal at 2500 volts and should you be running approximately 1 ampere of plate current you'll have to series-parallel eight of these just to produce 0.002 at 10 kV.

There you have it. Should you need further assistance, drop me a line (enclose an SASE).

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ham radio

VHF/UHF WORLD

Joe Reisert W1JR

medium power amplifiers

In one of my earlier columns I described VHF/UHF exciters, with the emphasis on transverters.[1] These transverters are primarily designed for 2 meters and 135 and 70 cm, but only at lower power — typically 0.5 to 1 watt output. In last January's and February's columns, I discussed high-power amplifiers, with the goal of attaining the legal power limit.[2,3]

But many VHFers need intermediate power level amplifiers, typically at 10 to 100 watts, for use as either a final stage or to drive a high-power amplifier. With this in mind, I've decided to dedicate this column to that subject.

amplifier types and classes

There are two major types of medium-power amplifiers: vacuum tube and solid-state. Each has its advantages and disadvantages. Generally speaking, tube amplifiers have higher gain, are more linear, are larger in size, and require at least two or more voltages. Solid-state amplifiers, on the other hand, are generally more compact and more physically rugged than tube-type amplifiers and usually require only a single voltage power supply.

Both types can usually be run in class C, which is all that's required for CW and FM opeation. However, nowadays most Amateurs prefer a linear amplifier, since it works equally well on CW and SSB at the flip of the mode switch and is less likely to cause key clicks when operating on CW.[1,2]

Medium-power vacuum tube amplifiers. Vacuum tube amplifiers have been around for a long time. They're usually quite reliable and rugged. Mismatches and mistuning — all too familiar to users of solid-state amplifiers — seldom cause catastrophic failures.

However, this type of amplifier can be bulky. It generally requires several voltages such as filament, plate, and possibly control and screen grid as well. And while it's not usually a problem, a warmup period of 1/2 to 5 minutes is usually necessary. Therefore, they're not particularly popular — especially for portable or mobile operation!

Regardless of their shortcomings, vacuum tube amplifiers are still quite plentiful and are often available at low prices. If you already have one anyway, why throw out or sell a perfectly good amplifier that's still functioning properly?

Many single-tube 2C39 amplifiers — often from the old Motorola T44 units — are still in operation on 70 cm. They make fine output amplifiers for low-power operation and serve well as drives from typical legal-limit amplifiers. There are also many single tube 4X150/4CX250B types of amplifiers still in use. They also make excellent moderate power amplifiers. Many modern high-power amplifiers use grounded grid circuitry and often have only 10-13 dB of gain. Hence they may require 75 to 150 watts of drive to achieve the full legal Amateur power limit. This is a great application for such an amplifier.

So don't dismiss the idea of using a vacuum tube amplifier as either a moderate output power amplifier or as a driver. (See references 2 and 3 for further information on the subject.)

Medium-power solid-state amplifiers. Solid-state amplifiers can be designed for either class C or linear operation. Both bipolar transistors and MOS power FETs are commonly used. There's no doubt that the compact solid-state amplifier has done wonders to increase VHF/UHF activity as well as reduce the size of the necessary equipment.

Power bipolars are presently the most common type of device used in Amateur medium-power solid-state amplifiers. They've been available for over 15 years and are moderately priced. Most work well at 12 volts DC and are therefore convenient for mobile operation. But they're usually quite fragile, and their transistors may burn out if subjected to a high VSWR load.

More recent power amplifier designs employ high-power MOSFETs, which became available about ten years ago. Manufactured by Siliconix, the first commercially available power FETs were called VMOS (Vertical Metal Oxide Semiconductors). Soon after, other manufacturers started making and improving on their performance, power, and reliability.

The original MOS power devices, primarily designed as switches, are generally called enhancement FETs, since they have to be biased "ON" to operate — just the opposite of the way vacuum tubes work. Because they had a high input impedance, they were subject to burnout when not connected in a circuit. Therefore, the manufacturers frequently placed a zener diode across the gate, thus reducing the frequency response and limiting use to HF and down.

Siliconix introduced the VMP-1 and VMP-4 devices in RF packages without the input protection zener diode. I've been using the same VMP-4 in a 10-watt 135-cm linear amplifier since

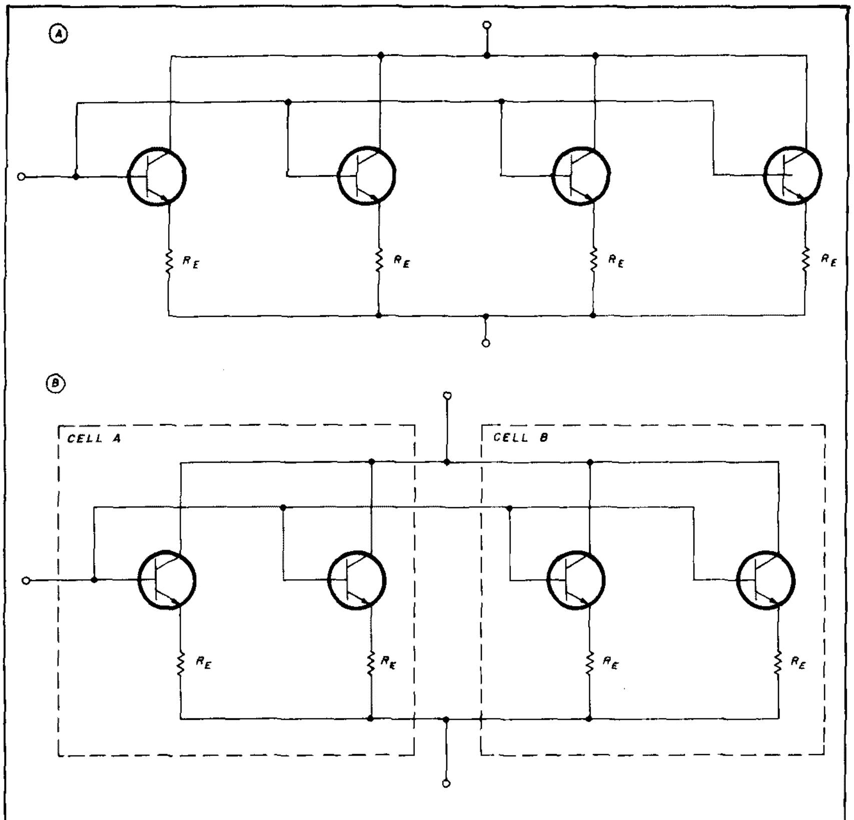

fig. 1. The typical bipolar power transistor is shown in schematic form. (*A*) single cell; (*B*) two-cell device. For values of R_E, see text.

1976! First it was my output stage and later my high-power amplifier driver.

Power MOSFETs are not as likely to be destroyed even if their output VSWR is high. When heated, many of these devices will just shut down rather than "self-destruct!" Power FETs generally have higher input/output impedances than bipolar power transistors, making them easier to impedance match (more on this later). However, the linear power MOSFET devices generally require higher operating voltage — 28 volts or more. For this reason, they're not as likely to be used in mobile or Amateur applications, where 12 volts is usually the only supply voltage available.

Power MOS devices stand an excellent chance of overtaking bipolar devices in many applications in the future. Although there may be some interest in the use of power MOS devices for Amateur applications, I'll limit this discussion to bipolar transistors because they're by far the most widely used solid-state devices in medium power amplifiers.

Before we can design and build a medium power amplifier, we must first know some of the basic properties of the bipolar transistor devices to be used. The most important parameters are the DC voltages, the power dissipation, recommended operating frequencies, RF input/output impedances and recommended circuitry for bipolar power devices.

The typical bipolar power transistor consists of many individual transistor junctions in parallel (**fig. 1A**). This configuration is chosen to increase current handling and distribute the heat within each junction. In the larger devices, there may even be multiple identical "cells" which are paralleled in the semiconductor itself to further increase the power and current handling (**fig. 1B**).

Resistance R_E (**fig. 1C**) is typically 5-30 ohms and is determined by the design requirements, manufacturing process, and material. The value chosen is very important, since the lower the resistance, the higher the gain of the transistor. This is frequently referred to as the ballasting resistor.

However, if heat increases in the junction, the typical bipolar transistor starts to draw more current and therefore heats up. If there's any potential difference between the junctions of the different transistors in parallel, the emitter current (and hence the collector current) may divide unevenly — and in an instant, there can be a chain reaction.

First one junction starts to "hog" current. It heats up until it's destroyed. Then the current is diverted to the remaining transistors, which may in turn be destroyed. The higher the resistance of R_E, the better the chances of equal current distribution (especially when heated) and the less possibility of (thermal) runaway destruction. Therefore the designer must carefully choose the optimum emitter resistance for the application and trade off gain versus power distribution and stability. Generally speaking, linear devices have higher R_E, while class C devices have lower R_E.

Power dissipation is a very serious consideration in the design of a power bipolar transistor. The junction area is quite small and the only way to cool it is with a good heat conducting package that is well heatsinked. This is in great contrast to power tubes where a large area is used and the power dissipation can usually be raised by increasing the air or water flow through the tube's plate radiator. Furthermore, the typical power bipolar transistor operates at a lower collector efficiency than its vacuum tube counterpart. While class C operation efficiency may approach 50 to 60 percent, typical linear service is usually between 33 and 50 percent.

DC voltages are most important.

However power bipolar manufacturers have made it easier for the user by optimizing the breakdown voltage and other parameters to suit the market. For example, there are power transistors for the 12 to 15, 24 to 30, and 50-volt markets. These markets are for the land mobile, aircraft/commerical and the pulse industries, respectively.

Current limits, also specified by the manufacturers, increase more or less in direct proportion to the power of the device and its application. In contrast to power tubes, the typical power bipolar transistor can be easily destroyed if the collector current rating is exceeded, even for an instant. Therefore, a current limited power supply is highly recommended. Fuses just may not operate fast enough to prevent burnout.

Most transistors are very frequency sensitive. For instance, the typical maximum gain of a bipolar power transistor usually decreases 6 dB every time the frequency is doubled. This means that if the transistor is operated at lower frequencies, it will have very high gain — typically enough to go into oscillation or "self destruct!"

Manufacturers work around this problem by designing each type of transistor to have a typical gain of 6 to 12 dB at the recommended operating frequency, much lower than that of a typical power tube! *Operation below the recommended frequency range is strongly discouraged.*

However, this is typically not a problem for Amateurs since most of us use a different power amplifier on each band. Also, power bipolar transistors are readily available for the Amateur 2-meter, 135 and 70-cm bands since they are also available for the commercial VHF, military and UHF bands, respectively.

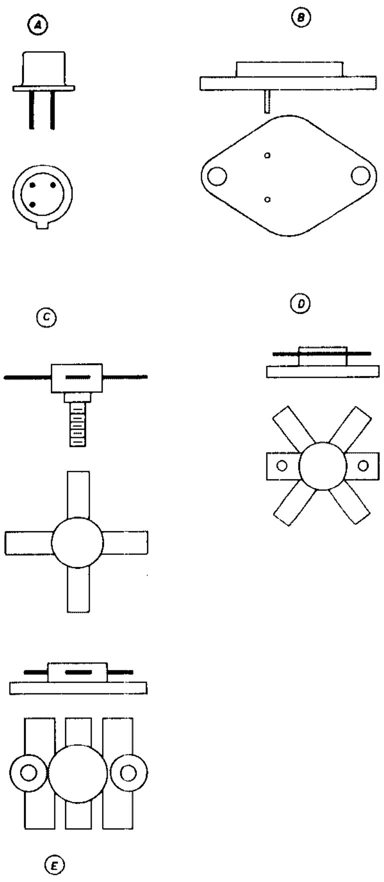

fig. 2. Typical bipolar power packages. (*A*) TO5/TO39; (*B*) TO-3; (*C*) stud-mounted; (*D*) high power; (*E*) high power, usually with internal matching.

RF input/output impedance. Unlike tubes, the input and output impedances of bipolar power transistors are extremely low and often have a significant reactive component. Input impeances of 1 to 10 ohms are quite common and the higher the power level, the lower the impedances will be. Typical input impedances are usually specified by the manufacturers either at midband or over the specified frequency band as plotted on a Smith chart. The output impedance of a power bipolar transistor is usually specified differently than the input impedance since it is a function of the internal device parameters, the operating power level, and the class of operation. Furthermore, the output or collector of a power bipolar transistor is usually not operated in a matched condition (often called "conjugate matched"). If the output were matched, this would mean that at least 50 percent of the power would have to be dissipated in the circuit, output efficiency and power would be lowered, and the device would have to dissipate more power. (This will be discussed shortly.)

Therefore, bipolar power transistors are often specified when operating into a *conjugate of the optimum load impedance* at a given frequency and power level. To simplify matters, most manufacturers usually show a typical or recommended circuit with the optimum components.

transistor configurations

Bipolar power transistors are usually operated in either a grounded emitter or grounded base configuration. Most modern devices operating below 500 MHz are specified for grounded emitter circuitry. Gain is usually slightly lower than in grounded base operation and the input impedance is higher, making it easier to match impedance. Grounded emitter configurations are also easier to bias and stabilize for linear operation.

Grounded base operation is still popular, especially above 500 MHz and where class C operation is used. Devices operating in grounded base circuits typically have higher gain (typically 1 to 3 dB), especially at the maximum frequency of operation. Another advantage is that the gain of a device increases only a moderate amount (10 to 15 dB) and flattens out as frequency decreases. Hence this circuit is less likely to oscillate at a lower frequency.

packages

Choosing the package for a high power bipolar transistor is almost as important to the device designer as designing the chip itself. At the higher power levels, heat dissipation is a complex problem. Because the area of the chip itself is usually quite small, manufacturers have selected packages with extremely low thermal resistance.

The typical three or four-lead TO-5 and TO-39 cans (fig. **2A**) cannot dissipate heat efficiently. TO-3 packages (fig. **2B**) are generally not suitable at VHF/UHF frequencies because of package parasitics.

Many power packages have been designed to handle power dissipation

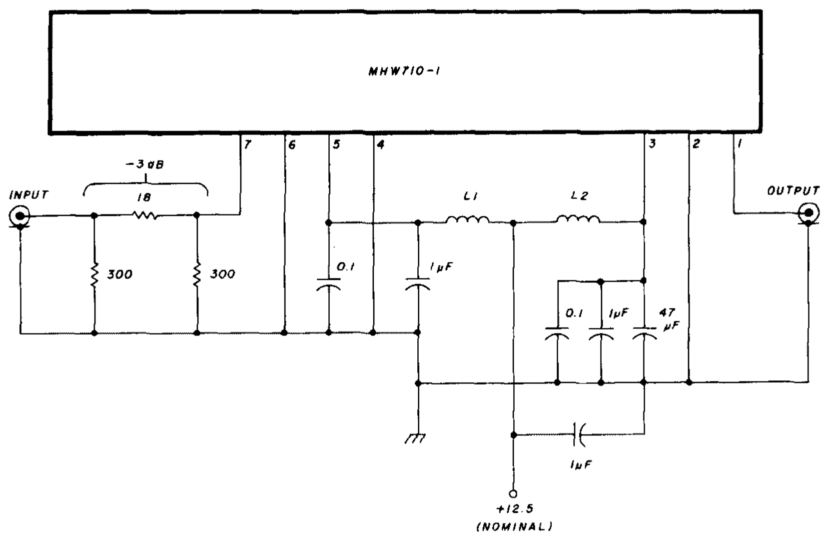

fig. 3. Hybrid power amplifier module. Unit shown is a Motorola MHW-710-1 for 400 to 450 MHz operation. Other suppliers and units may use slightly different pin layout.

while keeping the leads short. Furthermore, these packages usually have very wide leads to keep internal lead inductance at a minimum. At the lower power levels, 1 to 10 watts, stud-mounted packages are sometimes used (**fig. 2C**). However, at higher power levels, larger flat packages — generally with two mounting holes — are more common (**figs. 2D and 2E**). In all cases, the manufacturer uses a package commensurate with the power level of the device and specifies the thermal resistance and/or the type and size of heat sink required.

To further lower the internal parasitic load inductance, the common lead (either the base or emitter) is often attached to opposing leads on the package. Thus the package can be grounded on both sides. Not only does this improve gain but it often improves circuit stability.

Caution: Many UHF power transistors use a special low thermal resistance ceramic package containing beryllium oxide. If the package is crushed, ground, or abraded, the dust resulting from such action may be hazardous if inhaled. Therefore, never try to work with packages that may contain this material. To dispose of damaged or unwanted packages, enclose in an appropriate container for burial in an approved landfill, well away from ground water supplies.

linear versus class C operation

The first bipolar transistors were almost always operated in class C. To a great extent, this is still true today on FM. Class C design is rather straightforward and, as mentioned earlier, many devices are available with good gains.

However, most of the early devices available were not suitable for linear operation since they were highly nonlinear and tended to self-destruct if even a small biasing current was applied. Therefore, manufacturers typically redesigned their devices by adding higher ballasting resistors in the emitters, as discussed above. Careful attention to frequency response, stability, and gain lead to a whole new category of devices suitable for linear operation.

Nowadays most Amateurs use linear power amplifiers since they can be placed on CW, FM, or SSB. Ample information is available on class C operation, both in the literature and from manufacturers' data sheets. In addition, most suppliers have application notes available in their data manuals.

Before we leave class C operation, there is one type of commercially available unit that's particularly well suited for certain applications. Often referred to as the "brick" or "hybrid power amplifier module," this device is typically available at the 5 to 25 watt output power level for various frequency ranges from 144 to 940 MHz. They usually have high gain (2 or 3 internal stages) and therefore require only 0.1 to 1 watt of drive. Typically they are usable over 5 to 25 MHz of bandwidth (depending on operating frequency) with 30 to 50 percent overall efficiency.

Furthermore, these modules typically require a single 12 to 14 volt power supply and a few bypass capacitors. I particularly like them for driving frequency multipliers, which are more reliable if driven from a stable power source. Quite often I see these modules listed in surplus advertisements. Amperex, Motorola, RCA, NEC, and TRW are some of the major suppliers.

A typical 70-cm hybrid amplifier is shown in **fig. 3**. A good heat sink is required because efficiencies are low (30 to 40 percent). The 3 dB pad improves input match. However, hybrid modules are not suitable for linear service, and all the attempts to linerize them that I've heard of have been unsuccessful.

linear operation

Modern bipolar power transistors are still not capable of class A operation above a few watts because their power dissipation is too high for the packages and device geometries available. Most Amateurs prefer 12-volt power transistors because they're more compatible with the power supplies available (especially in mobile operation), moderately priced, and fairly rugged. However, if good linearity is required, the 24 to 28 volt devices are usually superior, even though they are less rugged and cost more.

When good linearity is required at power levels exceeding 5 watts, class

B biasing is preferred. Since linear devices are so sensitive to heat, special biasing techniques that "track" the heat within the transistor junction and readjust the bias voltage accordingly are required. Furthermore, these biasing techniques must be able to supply high base or emitter current on demand and from a low impedance DC bias source.

The first biasing circuits developed for bipolar linears consisted of a forward biased diode connected to ground and strapped to the package or heat sink as shown in **fig. 4A**. The series resistor to the power supply was adjusted until the proper collector current, usually designated as I_{CQ}, was attained. The power transistor idling current is typically 20 to 100 mA and is dependent on the device type and power level. It's usually not critical and often specified by the device manufacturer. Increasing the idling current rarely improves IMD, except on higher order products.

The idling current drawn by this biasing circuitry can be considerable, with 250 to 500 mA a typical value. In addition, this circuit is somewhat sensitive to power supply variations. Many designers, therefore, use 3-terminal voltage regulators in the supply circuit to keep the voltage constant (**fig. 4B**). The resistor is still used, but it dissipates less power (depending on the regulator voltage). A lower value resistor can be used with one of the newer low voltage adjustable regulators.

In 1973, Communications Transistor Corporation (CTC) introduced the "byistor," a self-contained regulator that incorporated temperature tracking (**fig. 4C**).[4] This device is actually a diode similar to the base emitter junction of a power transistor in series with a silicon resistor. The final DC source impedance is less than 1.0 ohm. Mounted in a studded package similar to medium power bipolar transistors, it can be attached to the heat sink, close to the transistor being regulated, for very close temperature tracking.

The byistor, like the diode biasing scheme, typically requires 300 to 350 mA of idle current. Some applications require low power drain. Therefore, in 1976 CTC introduced an alternative device, the Z0-28.[5] This device is actually a power bipolar transistor with two internal diodes (see **fig. 4D**). The diodes track the bias regulator and the transistor being regulated. The regulator transistor acts like a source follower. Hence this device has very low output impedance and draws high current only on demand.

To set up the Z0-28, a single low-power resistor is placed externally across the device as shown in **fig. 4D**. This resistor value is adjusted to set the power amplifier idling current. Some Amateurs have reported satisfactory operation by duplicating this device and circuit using discrete diodes (1N4001 types) and a power transistor. Close attention must be paid to adequate coupling to the heat sink to assure temperature tracking.

Other bias circuits have been used. When choosing a bias regulator, remember the following guidelines: the bias regulator circuit must track the temperature of the transistor being regulated and the output impedance of the bias source should be low, typically less than 1 ohm.

impedance matching

As mentioned earlier, the input and output impedances of bipolar power transistors are very low and often reactive. The typical "L" and "Pi" networks often used with vacuum tubes do not lend themselves to these impedances and devices, since losses may be significant.

The most frequently used matching schemes for high-power bipolar transistors are called "T" networks.[6,7,8] Several are shown in **fig. 5**. They are particularly adaptable if the input and output impedances of the amplifier itself are 50 ohms, the most common case. Each network has its advantages and disadvantages. Usually only one of the elements has to be varied. However, I prefer the schemes with two variable capacitors because they seem to be the easiest to tune and optimize properly, especially in narrow band

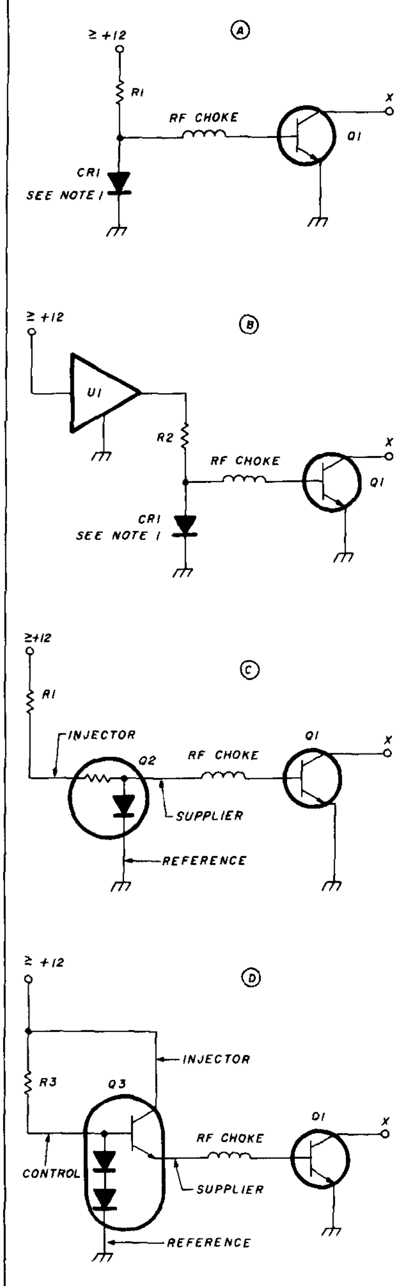

fig. 4. Typical linear biasing schemes. (*A*) single diode; (*B*) single diode with 3-terminal regulator U1 (see text); (*C*) Byistor, BY-1 (see text); (*D*) Z0-28 (see text). R1 and R2 are power resistors. Q1 is the power amplifier. Q2 is a BY-1 byistor and Q3 is a Z0-28. See text for other component value selections. Note 1: The biasing device must be thermally *attached* to the power device being controlled.

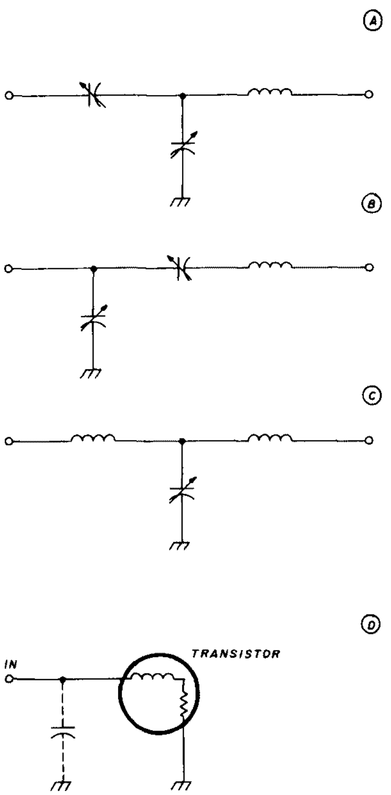

fig. 5. Typical impedance matching networks. (*A*), (*B*), and (*C*) are "T" networks. Value selection is discussed in text. (*D*) illustrates conversion from series to parallel equivalent circuit on transistor input as explained in text.

applications, and have a built-in DC block.

There is one rather sophisticated trick that is used extensively, especially in the *wider bandwidth amplifiers*. If the transistor series input impedance has an inductive reactance component (noted by a +j component such as the popular CM10-12A with an input impedance of 1.5 +j3.5 ohms), it can be mathematically converted from a series to an equivalent parallel network. Then an appropriate parallel capacitance can be placed at the input of the devices to tune out the inductive reactance as shown in **fig. 5D**. The transformed input impedance is higher and resistive. This makes the impedance matching network easier to design and with a smaller transformation ratio. Such techniques have been described elsewhere.[8]

This matching technique is also used extensively by the commercial suppliers of UHF transistors. First they adjust the series inductive reactance by adjusting the lengths of the bonding wires used to attach to the chip. Then they place the appropriate shunt capacitors internally in the package. The net result is more efficiency and greater bandwidth, as well as increased convenience for the circuit designer.

The output network is a function of the output power and the device. First the load impedance must be calculated using eq. **1**:

$$Rl = \frac{V_{cc}2}{2Po} \qquad (1)$$

where Rl is the desired output impedance, V_{cc} is the voltage across the transistor (usually the supply voltage less the saturation voltage of the transistor) and Po is the output power in watts. For example, with a 13-volt supply, a 1-volt saturation voltage (typical for most transistors), and a desired output power of 10 watts, Rl is approximately 7.2 ohms. This impedance is then converted in conjunction with the internal device impedance to the desired amplifier output impedance (usually 50 ohms).

Remember that the input network chosen is primarily matching the source (usually 50 ohms) to the power transistor. The output network is designed to yield the optimum load *impedance required for maximum* output power (as described earlier) with the required circuit loaded Q. If the proper component values and circuit Q are chosen, losses will be quite low and harmonics kept to acceptable levels.

Describing all the network requirements and design procedures is beyond the scope of this month's column. For those interested in the subject, I particularly recommend references 6 through 8. Many other papers have also been published. Most semiconductor manufacturers can offer applicable application notes.

Sometimes two power transistors are used, particularly when high power is required. These devices can be fed in parallel using a splitter network (**fig. 6A**) or with hybrid couplers (**figs. 6B** or **6C**).

Push-pull circuitry is often used at lower frequencies because it tends to cancel the second harmonics. Many modern transistors are now offered in matched pairs in a common package, making this technique very practical. Ninety-degree hybrids are also popular, since if one side of the amplifier fails, the output power drops to only about one-half. Another advantage of the 90-degree hybrid is that the amplifier input impedance match is quite good, since any mismatch is diverted to the external loads, R1 or R2.

recommended circuitry

The scope and size of this column do not permit me to describe a cookbook of circuits. Many suitable VHF/UHF power amplifier designs have been described both in Amateur and Commercial publications and on suppliers' data sheets. However, I will provide a universal circuit that can be used on the 2 meter, 135 or 70-cm bands (see **fig. 7**).

Note, in that circuit, that the input and output networks are similar to those just described. The diode bias network described earlier is also shown because it's very inexpensive; but the Byistor or Z0-28 devices (or their equivalents) are highly recommended as a replacement because they will definitely be more stable and reliable. *The variable capacitor and inductor values are only target values and can be varied slightly to obtain the desired performance.*

Since bipolar power transistors have much more gain below the operating frequency, they are often prone to self-oscillate at a lower frequency. This can usually be completely eliminated by the use of both low value and high value bypass capacitors as well as the network shown as L4 and R3. The low value capacitors should have short leads and be self-resonant well above the operating frequency. **R1** is not

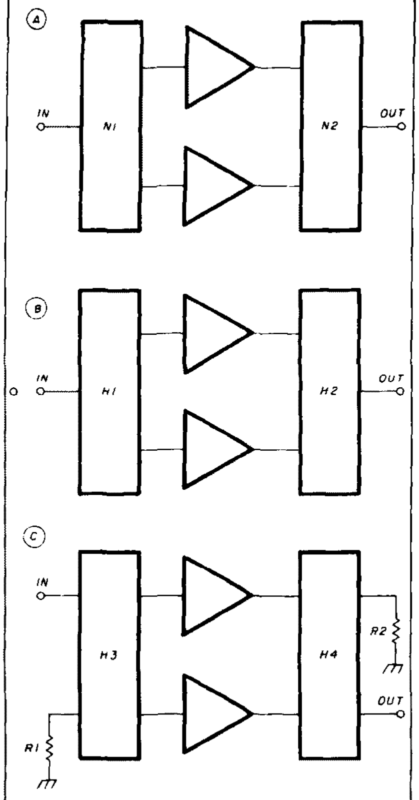

fig. 6. Various schemes for high power with combining networks. (*A*) in phase technique. N1 and N2 are reactive splitters; (*B*) H1 and H2 are 0 or 180-degree hybrid transformers. (*C*) H3 and H4 are 90-degree hybrid transformers. R1/R2 are 50-ohm loads. Resistor power rating depends on input or output power levels.

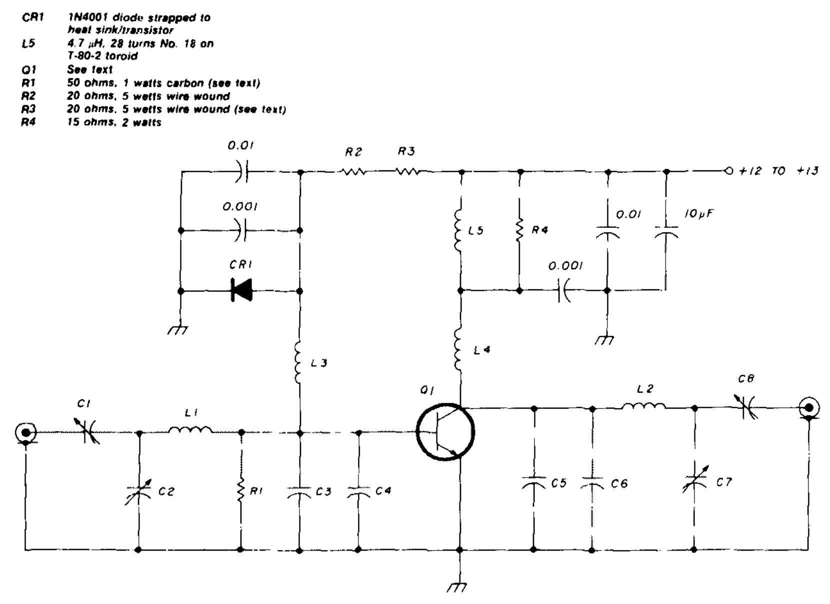

fig. 7. A typical power bipolar amplifier schematic with recommended starting values. See text for further explanation.

part of fig. 7

value	2 meters	135 cm	70 cm	units
C1	8-60	3-35	3-35	pF
C2	3-35	3-35	2-20	pF
C3	150 (Unelco)	100 (Unelco)	33 (Unelco)	pF
C4	150 (Unelco)	100 (Unelco)	22 (Unelco)	pF
C5	not used	100 (Unelco)	33 (Unelco)	pF
C6	not used	50 (Unelco)	33 (Unelco)	pF
C7	8-60	8-60	3-35	pF
C8	3-35	3-35	3-35	pF
L1	2 turns, No. 18, 0.25″ ID, 0.25″ long	2 turns, No. 18, 0.25″ ID, 0.25″ long	1.1″ × 3/16″ copper strap	
L2	2 turns, No. 14, 5/16″ ID, 0.25″ long	1 turn, No. 16, 0.25″ ID,	1.2″ × 3/16″ copper strap	
L3	0.33 RFC	0.33 RFC	0.1 RFC	microhenry
L4	4 turns, No. 18, 0.25″ ID, 0.5″ long	3 turns, No. 18, 0.25″ ID, 0.38″ long	2 turns, No. 18, 0.3″ ID, 0.3″ long	

always required, but often helps maintain stability in more stubborn cases. Since the input impedance of the power transistor is low, the value is not too critical.

component selection

The choice of components is very important. All capacitors should have short low-inductance leads and be able to handle the RF current often present when operating at these impedance and power levels. At VHF and low UHF, mica trimmers such as the Arco/Elmenco or equivalent are good for tuning elements. The Unelco type of sandwich micas are suggested for the collector RF choke bypass as well as in the fixed value input and output matching networks. These capacitors are easy to find at flea markets and are often available from surplus vendors.

At UHF and above, ATC (American Technical Ceramics) or equivalent porcelain chip capacitors are recommended. They have low loss, low inductance (if properly installed) and can handle high RF current without overheating.

Inductors should be large enough to keep the unloaded Q high.[9] The wire should also be large diameter, especially on the collector circuit, where, depending on power level, there can be high RF as well as DC amperes of current flowing.

You're probably wondering why I didn't mention microstrip or stripline inductors. These are fine if you're copying a working design, producing many copies of the same amplifier, or are proficient with the use of a Smith chart. However, for the typical one-

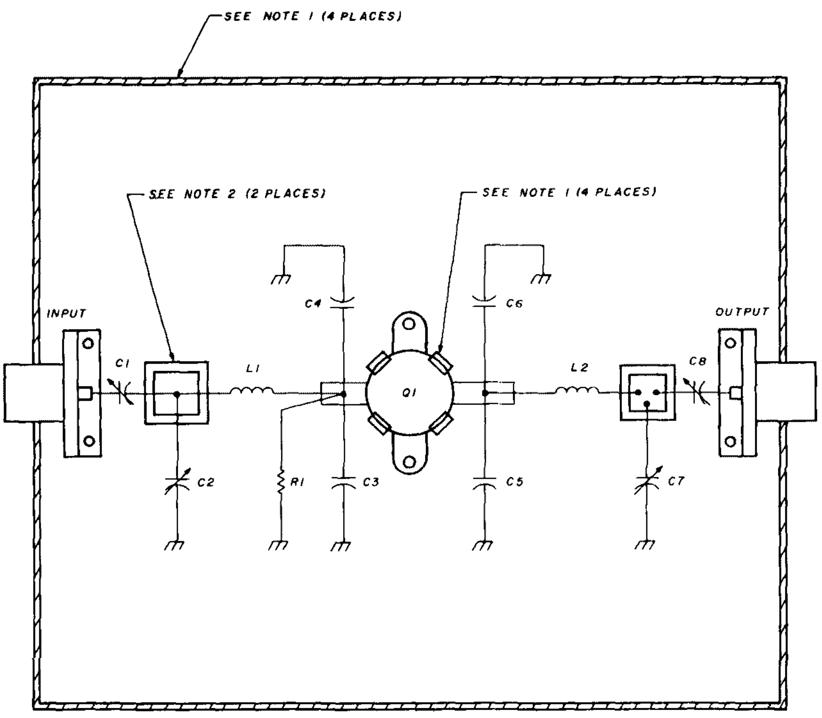

fig. 8. A typical construction technique that works well for VHF/UHF bipolar power amplifiers. Biasing and collector supply left off for clarity. Note 1: copper or brass foil soldered to top and bottom of PCB for low impedance ground. Note 2: Some PCB material has been removed to make an island to which components can be soldered.

shot design that is unproven, the discrete inductor is hard to beat. If you err in the value, just add a turn to or drop a turn from the coil and you're back in business. It isn't easy to cut up, extend, or decrease the length of a microstrip line on a printed circuit board if the tuning is incorrect!

In selecting transistors, choose those that will deliver sufficient power at the frequency of interest at the supply voltage available. As stated earlier, the 12-volt units are usually preferred for availability and price, but 28-volt units with better linearity would be a better choice, especially if you want to stay friendly with your neighborhood Amateurs. Consult manufacturers' data sheets. Don't be tempted to use a transistor at a lower frequency than recommended — it may self-destruct!

There are many power transistor manufacturers such as Acrian, Amperex, Motorola, RCA, Solid State Scientific, and TRW, to name a few. Order from the many suppliers who offer these devices in small quantities, manufacturers generally have a "minimum order" requirement that can make the per-unit price prohibitive. Many companies sell kits of parts, some even with a PC board. These are highly recommended for the Amateur who doesn't have a large junk box!

construction techniques

When building power transistor amplifiers, I prefer to build the circuits directly on or above a double-clad PC board type of material similar to the construction technique mentioned in reference 1. The choice of PCB material is not important because it's used only for ease of construction and to keep low impedance grounds.

For best performance, especially on UHF, the edges of the PCB should be wrapped with a thin copper or brass foil which can then be soldered to both the top and bottom of the board. This insures a good low-impedance ground. Next, drill a hole in the PCB sufficiently large to pass the power transistor package. Likewise, place a similar metal foil around this hole and solder it to the top and bottom of the PCB as shown in **fig. 8**. This will help keep the emitter (or base, if appropriate) at a low impedance to ground. Gain will not be reduced and the circuit will be more stable.

Then attach the PCB material to an adequately sized heat sink with an appropriate number of sheet metal or machine screws. The input/output connectors can be attached by right angle brackets at the ends of the board. Where appropriate, small squares or islands can be cut with a small sharp knife on the PCB as shown in **fig. 8**. Variable capacitors and other components can be tack soldered between these islands and grounded for mechanical circuit stability.

mismatches

Most modern power transistors are quite rugged, especially if they're emitter-ballasted. However, the voltages can soar if the amplifier looks into a high VSWR such as an open circuit. Therefore, always bring drive power up slowly and test for VSWR before running full power. If the VSWR is greater than 2:1, fix the problem before using your amplifier!

filtering

The harmonic content of a transistor power amplifier is usually quite high. This is why it's important to select the proper circuit operating Q and low-loss components as just discussed. Fortunately the amount of harmonic output acceptable (usually 40 dB minimum) is easy to obtain. Also your antenna system will often add some margin. If a transistor power amplifier drives a vacuum tube amplifier, the output harmonics will usually be lower due to the extra filtering usually present in tube amplifiers. However, if harmonics are unacceptable (as evidenced by TVI,

etc.), an output low-pass filter may be required.[9]

heatsinking tips

Several times in this month's column I've stressed heat dissipation as an extremely important parameter when using bipolar power transistors. The thermal resistance or conductivity to the heat sink must be low if long life and reliable operation are to be achieved. All power transistors should be directly attached to a heat sink without an intermediate surface. That is why I recommend the construction techniques above.

Thermal resistance can be kept low by applying a suitable heat sink compound to the transistor base where it mates with the heat sink. Use Dow Corning 340® or an equivalent compound. Mica washers or other objects should be avoided if at all possible.

The size and type of the heat sink is also important. Heat sinks with plenty of radiating fins are highly recommended. It's almost impossible to provide too much heat sink, but remember that the heat must travel *away* from the device, so the location of the attachment point is most important. Often the sink must be milled slightly to provide room for the nuts on stud packages. Don't remove any more fin area than necessary. (For those interested in choosing the correct heat sink, I strongly recommend that you refer to recent *ham radio* articles on thermal design.[10,11])

tuneup and test

At last we get to the bottom line — the final tuneup and operation. A recommended test setup is shown in **fig. 9.**

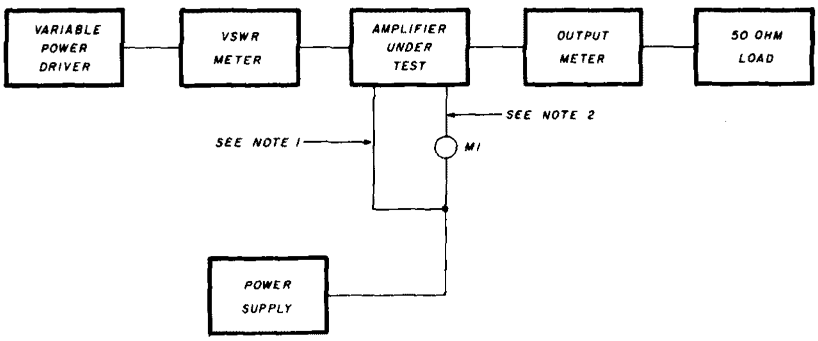

fig. 9. Test setup for tuning up a bipolar power amplifier. Note 1: to biasing circuits. Note 2: to collector circuit.

If the amplifier is operated in linear mode, first set up the I_{CQ} or collector operating current, as previously discussed. A 100 or 250 milliampere meter is usually sufficient. Temporarily break the collector line, making sure that the biasing circuitry is ahead of the meter. A variable supply is recommended for this step. Increase the supply voltage slowly while monitoring I_{CQ}. If the recommended current is obtained before the final operating voltage is reached, decrease the bias circuit current accordingly. Repeat this test and adjustments until the proper current is reached at the nominal operating voltage. If the transistor is stable and the heat sink adequate, the collector current should remain fairly stable.

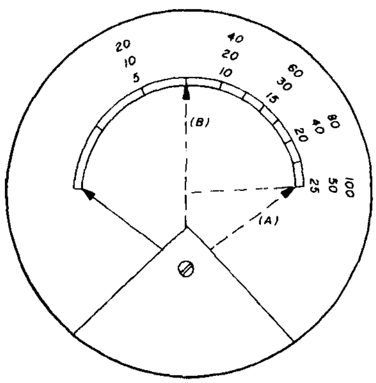

fig. 10. Typical RF power meter scale. For illustrative purposes, point A is the full power (key down) position. Point B is the maximum point at which the indicator should be when operating SSB for the same power output condition.

Next, short out the meter or put a high current type in its place. Apply a small amount of drive and quickly tune the output network for maximum output power. If none is forthcoming, tune the input network until the device starts to run power and adjust the output for maximum. Keep increasing input power while readjusting both the input and output tuning for maximum output power.

When the final desired or expected power is obtained, check the amplifier input VSWR. If it isn't 1.5:1 or better, adjust the input matching network accordingly. If you can measure collector current at maximum output power, calculate the amplifier's efficiency by dividing the indicated output power by the collector power (collector current times collector voltage). It should be at least 35 to 60 percent, depending on frequency and devices. If it is not, readjust the output tuning slightly to increase efficiency.

The final test is to see whether the amplifier output power is fairly linear. This can best be done by noting the amplifier gain (indicated power output divided by input power) at several power levels. At full power, the gain should be only slightly lower, perhaps 0.5 to 1 dB (80 to 90 percent), below that at lower power levels.

power meter syndrome

Let me broach one other subject before closing: I call it the "Power Meter Syndrome." Joe Ham tunes up to full power and sees 100 watts indicated on his wattmeter. Then he switches to SSB and proceeeds to watch the meter jump around. He's tempted to talk the power up to the same maximum tuneup level on the power meter, thinking that doing so will make him easier to copy at the

other end. Doing this causes splatter galore!

Why? Well, most power meters have some meter damping. What this means is that the indicator is slow to respond. Hence, on a voice peak, the meter will typically indicate only 25 to 30 percent of the peak power. I've illustrated this in **fig. 10**. Also, the maximum output point is probably at compression and therefore the IMD is poor at best. So, if you want to stay friendly with your local competition, keep your level down so that you're running only about 25 to 35 percent of the maximum possible output power as indicated by voice peaks.

conclusion

The intent of this article was to familiarize you with the bipolar power transistor and its use, rather than to provide a cookbook design approach. For some this will be sufficient. For others it will not be enough. However, with the material presented and the references provided, you should be able to forge ahead in the direction you choose.

Note: In several previous columns I've mentioned a home computer program called "RF-CAD." It's a very useful tool for designing filters and antennas and for matching, etc. This program is now available for use on the IBM PC or compatible machines. For further details and a list of capabilities, write Gary Field, WA1GRC, 5 Pluff Avenue, North Reading, Massachusetts 01864 (enclose SASE).

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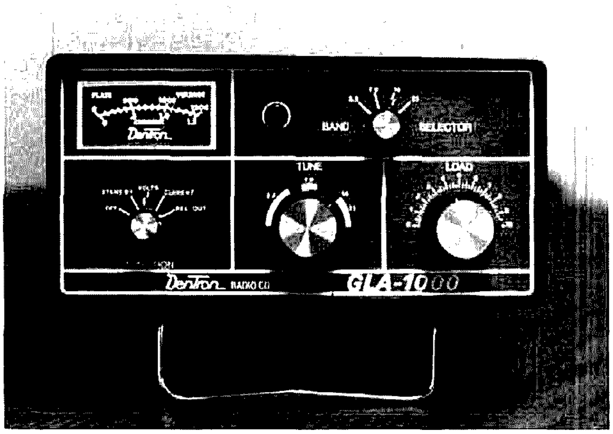
upcoming VHF/UHF events:

August 3-4:	*ARRL UHF Contest*
August 12:	*Predicted peak of Perseids meteor shower (0130 UTC)*
August 20:	*EME perigee*
September 7-8:	*International Region 1 VHF Contest*
September 14-15:	*ARRL VHF QSO Party*
September 16:	*EME perigee*
September 20-22:	*First annual 1296/2304 MHz Conference, Estes Park, Colorado (Contact W0PW).*

ham radio



Individual cathode bias does the trick

stop blowing finals
in the GLA-1000 amplifier

The Dentron GLA-1000 linear amplifier, introduced in 1978, has since become popular with HF operators on both sides of the Atlantic. Measuring only 11 × 5-3/4 × 11 inches (27 × 13.7 × 27 cm), this compact amplifier runs up to 1200 watts PEP input when driven by any of the popular 100-watt SSB transceivers. The amplifier uses four tubes in parallel operated in grounded grid configuration with 1200 volts on the anodes.

Though the tubes are capable of supplying the rated output with ease, problems were encountered often in daily use. The techniques used for solving the problem are applicable to all amplifiers with similar bias arrangements.

tubes destroyed with high SWR

My GLA-1000 linear amplifier operated well for the first three months until operation was attempted on 15 meters into a 3:1 SWR. The AC fuse blew and examination revealed that all the tubes were ruined. A replacement matched set of tubes was acquired from Dentron and installed, but suffered a similar fate when an accidental mismatch occurred.

In despair, I contacted Dentron for advice. They recommended updating the GLA-1000 into the latest GLA-1000B using a tuned circuit kit, a new grid bias zener diode, and add-on resistors for the anode parasitic chokes.

The original amplifier section of the unit is shown in skeleton form in **fig.** 1, which clearly depicts the bias arrangement.

tuned input kit

The modification kit replaces the fixed filter at the input with a PC board containing five pretuned pi circuits. These reduce the input SWR and improve the power transfer from the transceiver while filtering unwanted harmonics.

The zener diode in the modification kit is a 24-volt 1N3321 that replaces the original 9.1 volt 1N3308. This increases the negative grid bias, which reduces the no-signal anode current to 150 mA and reduces both the overall dissipation and the peak currents under drive conditions.

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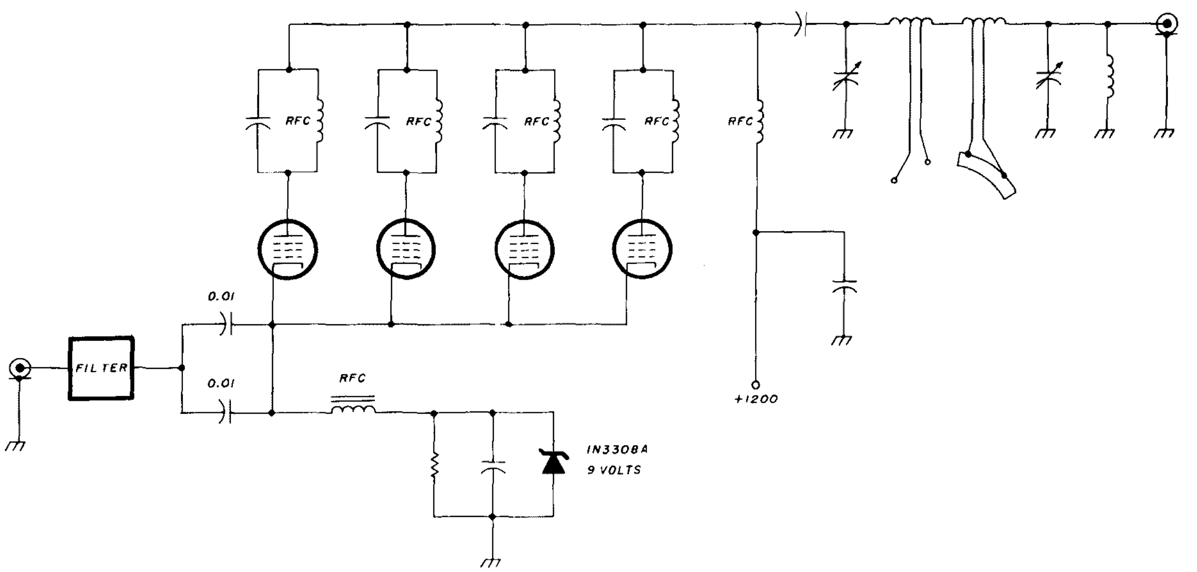

fig. 1. Simplified GLA-1000 original circuit.

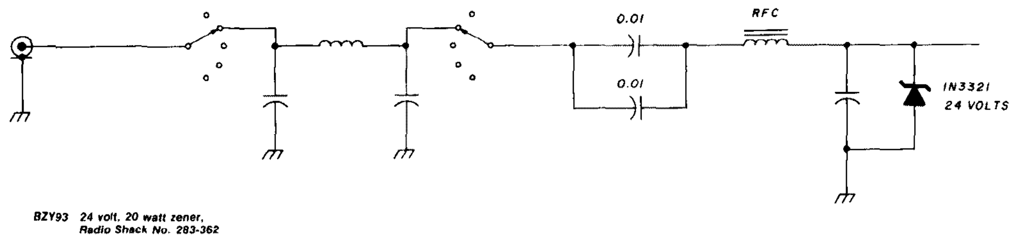

fig. 2. Dentron modification.

The four 100-ohm resistors added to the anode parasitic chokes are included to reduce their Q, which may prevent possible VHF instability.

A simplified diagram of the modified circuit is shown in **fig. 2**.

After modification, the third set of PA tubes lasted twelve months until an open circuit antenna feeder caused their demise together with the new zener diode. The circuit arrangement is far too critical of any mismatch, and instead of providing a fail-safe condition, it always destroys the costliest circuit components.

reason for sensitivity

Examination of the modified circuit reveals that the pi input circuit would not help power sharing between the tubes but would make the input look close to 50 ohms on all bands. While this is obviously an advantage to owners of all solid-state transceivers, it is of little help to operators of transceivers with tube power amplifier PAs that include their own pi output networks.

The zener diode replacement, obviously intended to control performance, does not accommodate unmatched sets of tubes. This is especially true if one tube is very different from the others. Under such circumstances the bad tube either shuts down by going gassy or shorts. In the shut-down case the remaining tubes would overdissipate and successively fail, just like dominos falling down. This is what appears to have happened the first two times. The third catastrophe would indicate a short circuit tube that blew the zener and left the remaining tube with no negative grid bias.

The bias is developed by drawing the cathode current of all four tubes from the single zener diode. This places the cathodes positive with respect to the grounded grids by the value of the zener voltage.

With the replacement of the Motorola zener diode likely to be expensive, drastic modifications to provide a separate bias supply were considered.

To control tube conditions both dynamically and quiescently, bias has to be supplied to each tube's cathode. This, together with the zener bias, would ensure equal power sharing. The idea was taken from the common practice used with parallel transistor stages, in which balancing resistors are used in the base or emitter circuit to overcome differences in base/emitter voltages and to ensure equal power sharing. **Figures 3** and **4** show typical transistor power-sharing circuits.

Automatic bias is provided simply by adding a resistor in the cathode so that cathode current produces grid bias. The value of resistor has to be chosen with care so that it does not drop too many volts on current peaks and cause flat topping. Nevertheless, it has to be large enough to cause about 1 volt change of bias for a 50 percent change of cathode current. With a combined current of 150 mA, each tube should be drawing 37.5 mA, about right for most class AB1 tube finals. A 50 percent change would be 18.75 mA, so for 1 volt the automatic bias resistor R_X would be

$$R_X = \frac{1V}{18.75\ mA} = 53.33\ ohms,$$

(a 56 ohm resistor was chosen).

However, since the signal is also applied to the cathodes in this configuration, the 56-ohm resistors in series would cause appreciable loss of drive and also mismatch the newly installed pi input circuit. To over-

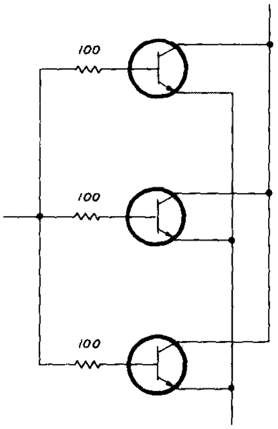

fig. 3. Bias and drive balancing.

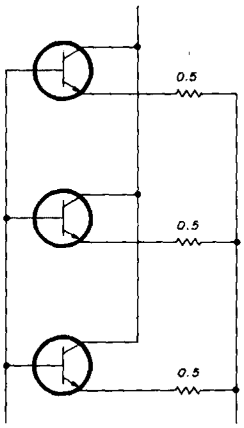

fig. 4. Bias and load sharing.

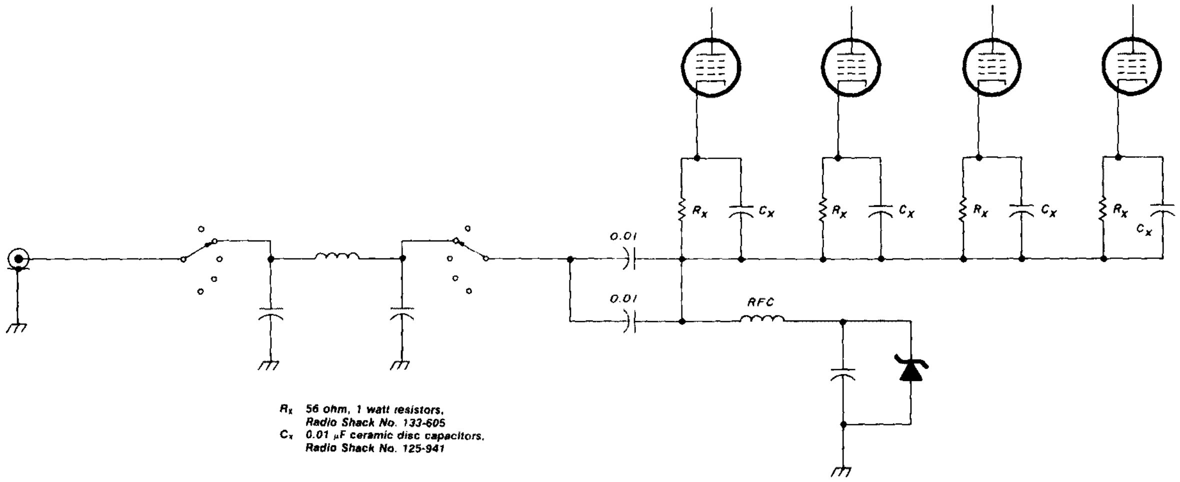

fig. 5. Bias and load sharing modification.

come this, each resistor could be bypassed to RF by a low reactance capacitor. A value was chosen by trying standard values in the formula $X_c = \frac{1}{2\pi fC}$ where it was found that 0.01 μF was 4.55 ohms at 3.5 MHz and 0.57 ohm at 28 MHz.

The new circuit configuration is shown in **fig. 5**.

construction details

The tube sockets in the GLA-1000 are mounted on a PC board. A scale drawing of this is shown in **fig. 6**. The 0.01μF bypass capacitors, C_X, are attached to the resistors as shown and wired across the breaks marked on the tracks connecting pins 3 of the tubes.

A suitable alternative to the 1N3321 zener is the BZY93 24-volt type. Available at low cost, it has a stud anode instead of cathode. It can be fitted into the hole used by the old diode but must use an insulating kit and solder tag to pick up the positive stud. Use a grounding strap on its wire end.

performance

It was well over a year since the modification was done. Since that time, the unit has undergone intense activity and repetitive abuse. The absence of TVI complaints suggests that the modification has not degraded signal purity in any way. The peak output power is easily achieved and appears stable under continued full power operation.

The current sharing technique prevents the "domino effect" breakdown under fault conditions. This was convincingly demonstrated when a severe mismatch caused the AC fuse to blow — but nothing else was damaged!

The amplifier was designed to use selected and matched tubes designated D50 by Dentron. Nevertheless the third and fourth sets were standard type 6LQ6 manufactured in the USA. Tube types 6LQ6, 6JE6, or 6MJ6 may be used, although idling currents may differ sufficiently to require adjustment of the bias zener voltage to give between 35 and 40 mA quiescent current. The 6LQ6 tube and the 6MJ6 type, which are capable of higher power at lower frequencies but unsuitable for 10-meter use, are both readily available. Although Japanese tube substitutions were found to be unstable in this design, both before and after modification, they appear to provide higher gain. The American types are the lowest cost and have shown no signs of instability. (Higher power types 6MJ6 have not been tried at any time.)

This modification has been so successful that I've shelved my half-built homebrew 813 linear. Undoubtedly the principles involved can be extended to other linears of similar design and should result in greatly extended tube life.

ham radio

a pulsed, constant current, NiCad battery charger

Longer life, shorter charge time — plus automatic shutoff

"Zapping" **NiCad** batteries with a burst of high current is an old trick often used to revive seemingly dead cells. Word has it that some RC model airplane enthusiasts extended this idea to the actual charging of batteries, with favorable results; their batteries lasted twice as long and charged more quickly than those who used conventional charging methods. The idea was to charge with a train of pulses instead of a steady current. Apparently, less energy was lost to heat during the charging process, resulting in shorter charging times and increased battery life.

This charger was designed with these goals in mind. Over the two years it's been in use, charging times have been shortened by one-third and new life has been given to a battery pack about to be thrown away.

principle of operation

The key to pulse charging is to keep the average value of the charging current the same as it would be in conventional schemes, but use a low duty cycle waveform. The plot shown in **fig. 1** is the current waveform used in this charger. For 10 percent of each cycle, current equal to I_{MAX} charges the battery. For the remaining 90 percent of the cycle, no current flows. This produces a rectangular waveform with a 10 percent duty cycle.

The DC, or average value of this waveform is 10 percent of the pulse amplitude, or

$$I_{DC} = 0.1 I_{MAX} \tag{1}$$

As long as normal charging precautions are observed and the DC charge rate is within the capability of the battery, pulse charging will in no way harm the battery. A pulse width of about 100 microseconds (μs) is appropriate for this application, which, because of the 10 percent duty cycle, yields a pulse rate of 1 kHz.

circuit description

The block diagram in **fig. 2** outlines the basic design of the charger. The schematic is shown in **fig. 3**. The charger — designed to handle from one to ten NiCad cells, with an adjustable charge rate of 50 to 450 mA — consists of the following sections:

Constant-current source. The charging current is supplied by a constant-current source that can be pulsed at peak currents of up to 5.0 amperes. A constant-current source can be made from a three-terminal voltage regulator and a resistor, as shown in **fig. 4**. In this circuit the current, I_{DC}, is simply the output voltage of the regulator, V_0, divided by R3, or

$$I_C = \frac{V_0}{R3} \tag{2}$$

Unfortunately, this method does not easily lend itself to high frequency switching.

A better approach is shown in **fig. 5A**, in which the regulator is replaced with a zener diode and pass transistor. Here, the current is equal to the zener voltage, V_Z (less the base-emitter voltage drop of Q1) divided by R3, or

$$I_C = \frac{(V_Z - 0.6)}{R3} \tag{3}$$

The current can be instantly switched off by grounding the base of Q1. This circuit worked well except at high currents.

To improve its operation at high currents, I added a second transistor as shown in **fig. 5B**.[1] This circuit worked very well. It could supply a constant current into either a short circuit or 10 cells in series and still hold the current to within 1 percent of its setting.

As shown in **fig. 3**, the current is adjusted by R3. R4 is inserted in series with R3 to limit the current if R3 is set to zero.

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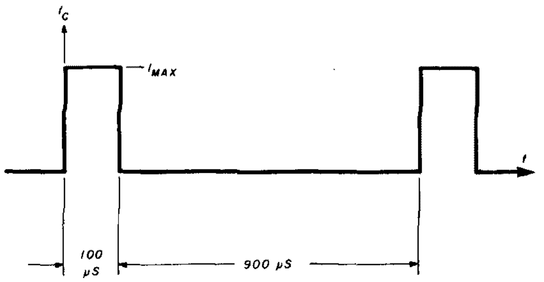

fig. 1. The charging current has a rectangular waveform with a 10 percent duty cycle. Repetition rate is approximately 1 kHz.

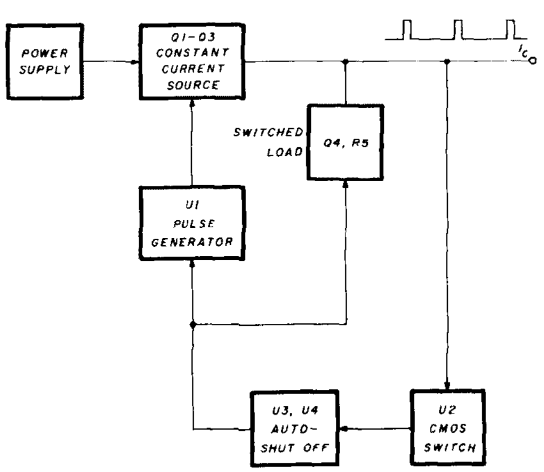

fig. 2. Block diagram of the pulse charger.

Power supply. For the constant current circuit to work properly while charging up to 10 cells in series, the power supply must provide at least 30 VDC. The unregulated power supply shown in fig. 3 satisfies that requirement if T1's taps are wired for 22.3 VAC, as shown. A separate zener regulated section provides 15 VDC for the control circuitry of the charger.

Pulse generator. The pulse waveform is generated by U1. Actually, U1 produces a 90 percent duty cycle waveform instead of 10 percent in order to accommodate the inverting action of Q3. Q3 turns the current source off when the output of U1 is high, and on when it is low. That converts U1's 90 percent duty cycle to a 10 percent duty cycle charging current.

Automatic shutoff. Different schemes abound for controlling the charging time, each with their advantages and disadvantages. In the scheme chosen for this charger, charging stops when the battery reaches a set voltage. However, the battery voltage is not monitored in an ordinary manner.

Because of a battery's internal resistance, the voltage at its terminals during charging is less than its open circuit voltage, and will vary with the charge rate. As soon as a load is put across the battery, the voltage will drop further. In this charger, battery voltage is monitored only during that part of each cycle in which the charge current is off. Furthermore, a *small load*, R5, is kept across the battery to give a truer indication of the state of charge when the battery is monitored. The load is automatically switched out of the circuit during automatic shutdown.

Automatic shutoff is accomplished by monitoring the battery voltage with comparator U3. R15 and R16 divide down that voltage and U3 compares it with the voltage provided by reference diode CR8. When the voltage at the comparator rises above the reference voltage, Q5 turns on, shorts out C5, and stops U1 from oscillating. That forces the output of U1 high, which holds the constant-current source off.

R13 and R17 add hysteresis to the action of the comparator. With the values given, the battery voltage must drop about 30 percent before charging will switch on again after automatic shutoff. The amount of hysteresis can be changed by raising or lowering the value of R17.

In order for U3 to monitor the battery voltage only during the off portion of a charge cycle, CMOS switch U2 passes the voltage on to the comparator in sync with U1. In this way, when U1 stops oscillating, the comparator will still see the battery's voltage because U2 is kept closed by U1 in the automatic shutoff state.

Current metering. The torque on the pointer in a standard D'Arsonval meter movement is directly proportional to the true DC value of any current passing through it. As a result, the DC milliamp meter M1 in **fig. 3** correctly indicates the true charge rate delivered to the battery. Load resistor R5 is wired in the circuit so that the load current delivered to R5 by the battery during the off period of a cycle is taken into consideration by M1. The value of R5 was chosen to add a load of about 25 mA when charging an 11-volt battery pack.

construction

Perfboard was used for mounting most of the components. The board was connected to the rest of the circuit by an edge connector to make it removable during the design phase of the project, but any convenient construction method can be used because parts layout is not critical. A Radio Shack cabinet (former catalog No. 270-269, fig. 6) housed the circuit. M1 is specified as 0-500 mA, in the parts list; a 0-100 μA movement was used in the original model by adding a current shunt.

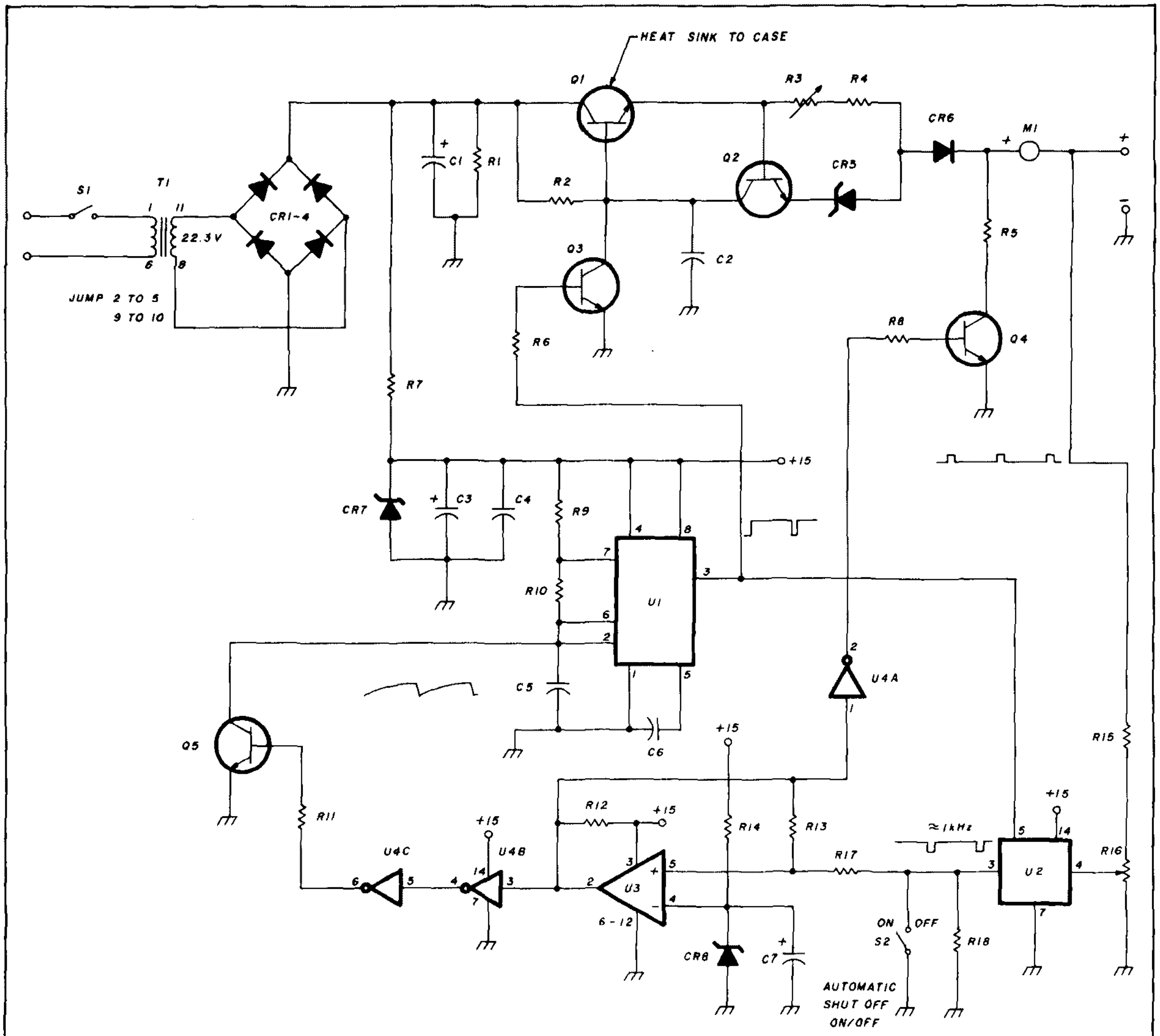

fig. 3. Schematic of pulse charger. U2-U4 can be eliminated if the automatic shutoff feature is not wanted.

Although pass transistor Q1 is heat-sinked to the case, it must be electrically insulated from it. During conduction, the current through Q1 can be as high as 5 amperes and the voltage across it about 15 volts. Instantaneous power is $5 \times 15 = 75$ watts, but the power which must be dissipated by Q1 is only ten percent of that because of the duty cycle. Using the case as a heat sink works well for this low power level.

The heat dissipated in R3 and R4 is proportional to the square of the RMS voltage across them, divided by their resistance. Because of the ten percent duty cycle, that RMS voltage is approximately one-third of the peak voltage across the resistors, as is the RMS current value through the resistors. With this in mind, the wattage ratings should be 10 watts for R4 and 5 watts for R3. An old loudspeaker level control was used for R3, but any 70-100 ohm, 5-watt potentiometer will do. Since the charge current is inversely proportional to R3, the higher current settings will bunch up at one end of the shaft rotation. A potentiometer with a non-linear taper can reduce that effect.

Switch S2 disables the automatic shutoff feature. If desired, the entire shutoff feature can be left out by removing U2 through U4 and their related components, as well as R5 and Q4.

operating the charger

NiCad batteries can be charged at a significantly higher rate than that provided by the average stock charger as long as the battery is not overcharged.[2] Once fully charged, the battery must dissipate any additional charging as heat. At the popular C/10

Parts list figure 3.

item	description
C1	1500 µF, 50 volt electrolytic
C2	0.001 µF
C3	100 µF, 25 volt electrolytic
C4,C6	0.01 µF
C5	0.01 µF mylar, 10 percent
C7	10 µF tantalum
CR1,CR2	
CR3,CR4	6 ampere, 100 volt rectifier bridge
CR5	1N753 6.2 volt zener diode
CR6	2 ampere, 50 volt rectifier diode
CR7	1N4744, 15 volt zener diode
CR8	1.2 reference diode
M1	0-500 mA DC ammeter
Q1	TIP120 (Radio Shack No. 276-2068)
Q2	TIS97
Q3	2N3904
Q4	2N2222
Q5	2N3704
R1	2.2 kilohms, 1 watt
R2	2.2 kilohms, 2 watts
R3	70-100 ohms, 5 watt potentiometer (see text)
R4	1.5 ohms, 10 watts
R5	470 ohms, 2 watts
R6	1.5 kilohms
R7	680 ohm, 2 watts
R8,R11	10 kilohms
R9	120 kilohms
R10,R18	22 kilohms
R12	3.3 kilohms
R13	2.2 megohms
R14	15 kilohms
R15	5.6 kilohms
R16	2 kilohms, 10-turn trim potentiometer
R17	51 kilohms
S1,S2	SPST toggle switch
T1	Stancor RT-201 power transformer or equivalent
U1	NE555 timer
U2	CD4016 CMOS switch
U3	LM339 comparator
U4	National CMOS 74C04

All other capacitors are disc ceramic. Except as noted, all resistors are 1/4 watt.

charge rate, where C is the ampere-hour capacity of the battery, NiCad cells can be overcharged continuously without risk of damage from overheating. This assures safety in case the battery is left on "charge" continuously. On the other hand, if the battery is charged at a higher charge rate, say C/3, and goes into overcharge, cell temperature will rise, with possible damage and loss of battery life. However, if care is taken to avoid overcharging, there is no inherent reason why a NiCad cell cannot be charged at the higher rate. The automatic shutoff will prevent just such overcharging from occurring after it is properly set.

To set the automatic shutoff, first remove any diode in series with the HT's charging jack if the battery will remain inside the HT during charging. (If you don't do this, the automatic shutoff feature won't work.) Connect the discharged battery to the charger, turn the charger on, and turn off the automatic shutoff with S2. Set the charge rate to C/3, and add another 5 mA to compensate for changes in the load of R5 as the battery voltage rises. (For a 500 mA hour battery, that would be about 155 mA.) Monitor the battery voltage with a precision voltmeter, preferably a DVM — accuracy is not as important as precision. Check the

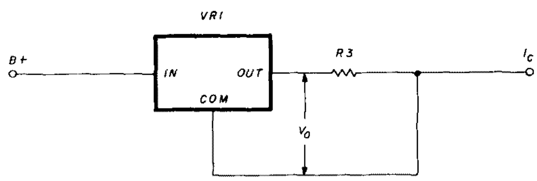

fig. 4. A 3-terminal regulator forms a constant current source by holding the voltage across and, hence, the current through R3 constant.

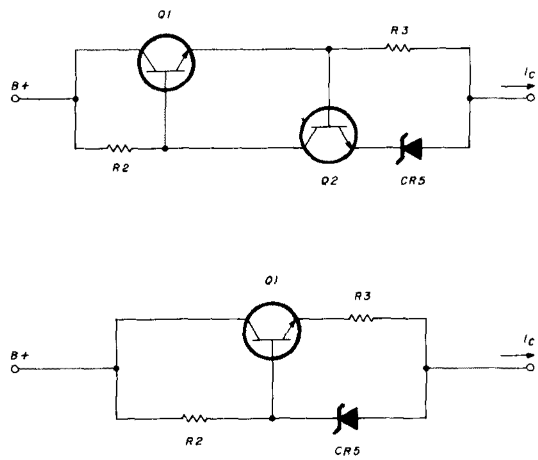

fig. 5. A constant current source can be created with a zener diode, one or two transistors, and a couple of resistors. The current can easily be switched off to form pulses by grounding the base of Q1 through another transistor.

battery frequently as it charges. The battery voltage will rise rapidly initially and gradually thereafter. When the battery nears its fully charged state the voltage will again rise, but more importantly, the battery will start to become warm to the touch. This increase in temperature indicates that the battery is now fully charged. Note the voltage when this occurs; it will probably be the equivalent of 1.45-1.50 volts per cell. Also make a note of how much charging time it took to arrive at this point, starting with a fully discharged battery.

Now turn on the automatic feature, and adjust R16 until the charge current turns off. Reduce the charge rate setting slightly and reset the automatic shutoff by cycling S2 on and off. Raise the charge current back up again and watch the voltmeter to make sure charging shuts off at the voltage just measured. If the shutoff point cannot be set exactly, err on the low side by setting it slightly below the desired voltage.

If the automatic feature is not used, charging can

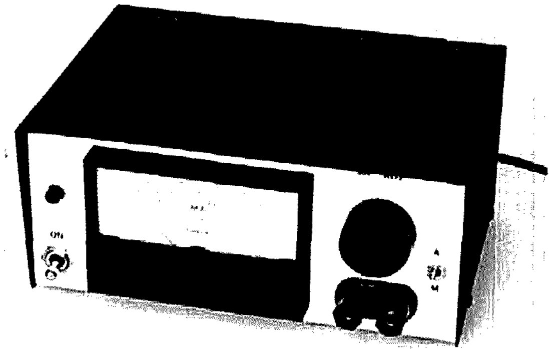

fig. 6. The charger fits nicely in a standard Radio Shack cabinet. Parts layout is not critical and any convenient construction method can be used.

be controlled by time. An inexpensive mechanical timer can shut off the charger after the proper length of time, as just determined. Remove the "ON" arm of the timer to keep it from again turning on the charger 24 hours later.

conclusion

The pulse charger has been a considerable time-saver: it takes only 3-1/2 hours to fully charge a 500-mA hour battery pack at a 150 mA charge rate; a non-pulsed charger at that same rate took 4-1/2 hours. This represents a time savings of over 30 percent, and increased charging efficiency as well. As a test, the battery was discharged with a load resistor equivalent to a 100 mA load and timed to confirm that it was indeed fully charged by the pulse charger.

The charger was also used to add life to an old, seemingly dead battery pack. The battery was consecutively discharged and charged about four times. From that point on, it lasted another three months.

The pulsed charger should prove to be a worthwhile accessory for anyone who uses NiCad batteries extensively.

acknowledgement

Many thanks to Pat Spadafore, KA2MOV, for his invaluable experience and recommendations concerning pulsed charging techniques.

references

1. *The Radio Amateur's Handbook*, American Radio Relay League, Newington, Connecticut, 1983, page 4-31.
2. *Nickel-Cadmium Battery Handbook*, 2nd Edition, General Electric Company, 1975.

ham radio

3CX1200A7 10 to 80-meter amplifier

New EIMAC tube provides 1500 watts out

For several years, I have used the popular EIMAC 3-1000Z as the main amplifier tube in my station. I enjoy the instant-on characteristics of the tube as well as the well-matched drive requirements for a 100-watt output exciter. Recently EIMAC introduced the 3CX1200A7, the newest addition to its line of metal ceramic external anode triodes. The electrical characteristics of the tube are identical to the 3-1000Z except for the increased plate dissipation (1200 watts) allowed by the external anode.

I decided to modify my 3-1000Z RF deck to accommodate the new tube. The only change necessary was in the tube socket and chimney. (The 3CX1200A7 uses the EIMAC SK-410 or Johnson 248 socket commonly used for the 3-500Z tube. A matching glass chimney — EIMAC SK-436 — is also available). The modification took less than eight hours. This article summarizes the results of the project.

the amplifier circuit

As indicated above, this amplifier is designed around the new 3CX1200A7 (see **fig. 1**), using a tuned input network ganged to the main bandswitch. The input circuit minimizes distortion products and helps provide a 52-ohm match to the exciter. Maximum SWR presented to the exciter is 1.2:1. Approximately 100 watts is required to drive the amplifier to 1500 watts output. An effective ALC circuit is also included to prevent overdrive by higher power exciters.

Output impedance matching is accomplished with a pi-network designed for a Q of 12. Toroids are used for compactness for the 80-meter pi coil and the tuned input network coils.

Also included is a grid trip circuit to disengage the amplifier should the grid current exceed 300 milliamperes. This circuit protects the tube in case of excessive drive or an improper load presented to the amplifier.

The amplifier includes an effective dynamic bias circuit. Response time of the circuit is fast enough to cut the amplifier off between syllables on SSB and between dots and dashes on CW. A defeat switch is included on the rear panel of the amplifier. When engaged, the heat generated by the tube is substantially *reduced* during amplifier operation as feeling the temperature of the air exiting the amplifier cabinet will confirm.

A 100 VDC power supply is included to power the dynamic bias circuit, and a 26 VDC supply is included for panel lights and relay operation. An RF wattmeter is included for tuneup and measurement of amplifier efficiency.

amplifier control circuitry

The amplifier control circuitry is illustrated in **fig. 2**. 120 VAC enters the RF deck from the power supply. Power is applied to the amplifier by depressing the front panel switch S1. S1A applies 120 VAC to the HV power supply, the time delay relay TD1, the filament transformer, the 100 VDC bias supply and the 26 VDC supply. The blower receives power immediately through S1B. When the power is turned off with S1, the blower (*B*) keeps running for approximately three minutes to cool the 3CX1200A7. This delay is accomplished through the use of the time delay relay TD1. The control circuitry is quite simple because of the "instant on" characteristics of the 3CX1200A7 tube.

grid trip protection

A grid trip module has been included in the amplifier design to protect against high levels of grid current which could be dangerous to the 3CX1200A7 tube. High levels of grid current could be caused by excessive drive, improper tuning or lack of a 50-ohm load on the output of the amplifier. In this design, if the grid current exceeds 280 milliamperes, the circuit trips relay RL3 (see **fig. 3**), which breaks the VOX amplifier line to deactivate the amplifier and lights the front panel "grid trip reset" push button (S4). It is necessary to push the reset button to put the amplifier back into operation. Of course, one should determine why the amplifier exceeded 280 mA before proceeding with amplifier operation.

The circuit operates as follows: grid current is drawn through the 10-ohm resistor (R1) as shown in **fig. 1**. The current flowing through the 10-ohm resistor develops a voltage that turns on transistor Q3 in the grid trip circuit when the current reaches 280 mA (or any other value if desired). 280 mA through a 10-ohm

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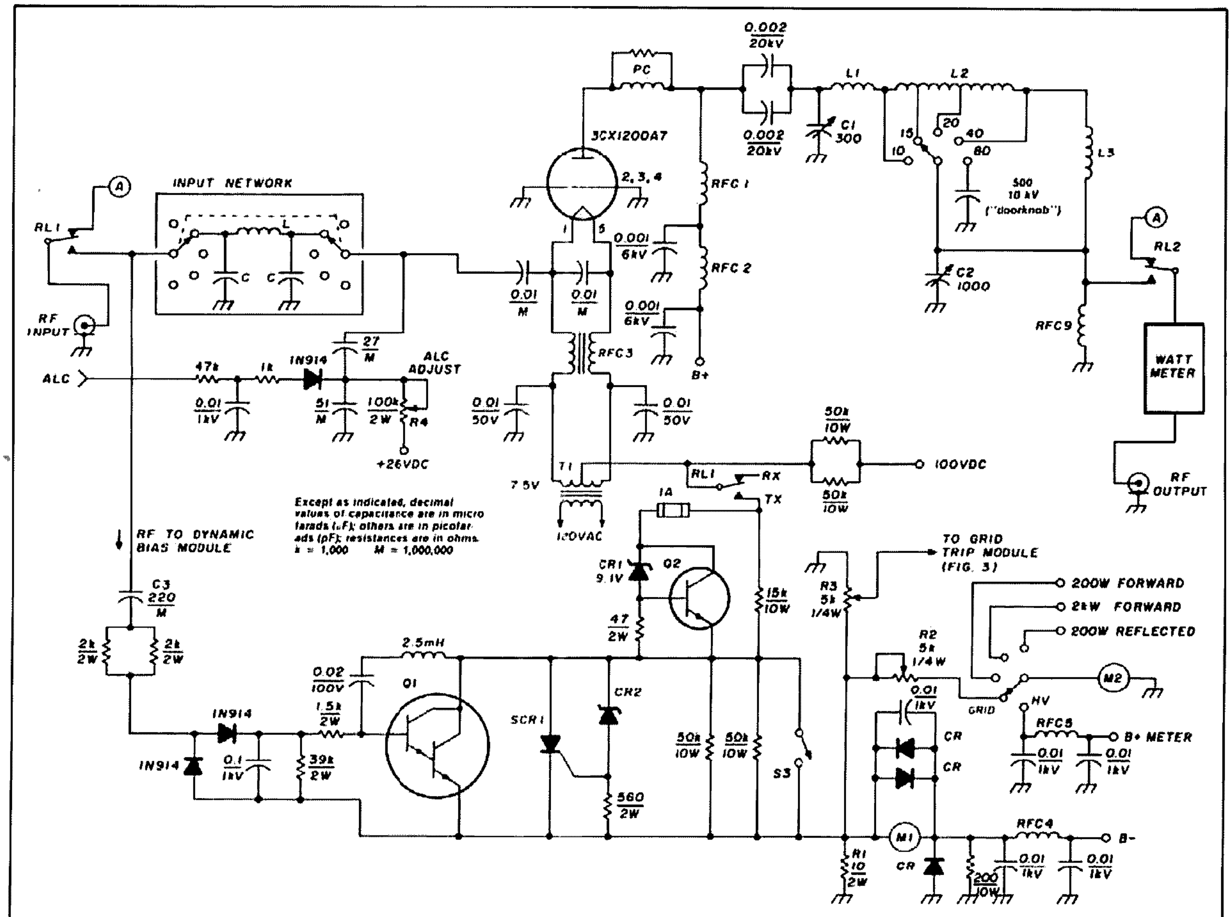


fig. 1. 3CX1200A7 RF deck schematic.

3CX1200A7 amplifier provides full legal output on 10-80 meters.

resistor results in the development of 2.8 volts. Only 0.6 volts is necessary to turn on Q3. Therefore, a 5-kilohm trim pot (R3) has been included as a voltage divider for adjustment.

When Q3 turns on, it serves as a switch providing a path to ground to actuate relay RL3. A set of the relay contacts physically grounds the relay coil of RL3, taking the current load off Q3. If this feature were not provided, the transistor would start gating the amplifier on and off. It is therefore essential to latch the grid trip relay closed. Another set of contacts on relay RL3 breaks the VOX line, deactivating the amplifier. A third set of contacts applies power to the pilot light on the front panel "grid trip reset" pushbutton (S4) located on the front panel of the amplifier to make the operator aware of what happened. Pushing S4 breaks the path to ground for relay RL3, deactivating the relay and putting the amplifier back into a ready state.

This circuit provides simple and effective protection from costly mistakes during amplifier operation.

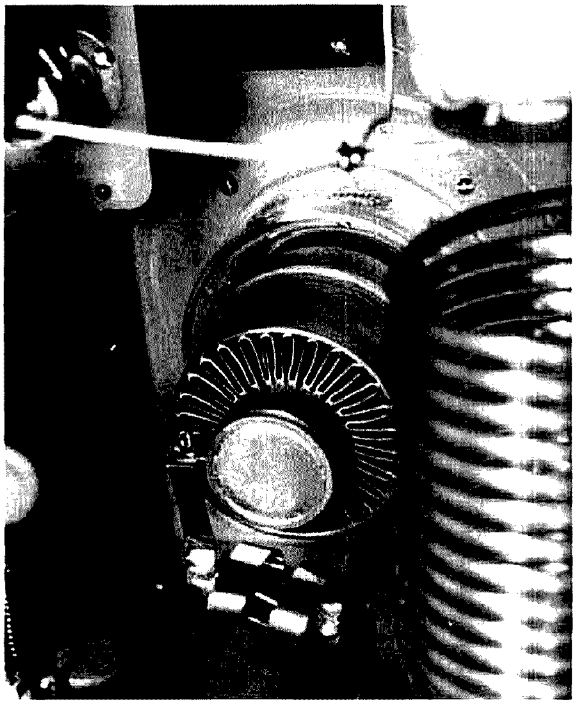
dynamic bias circuit

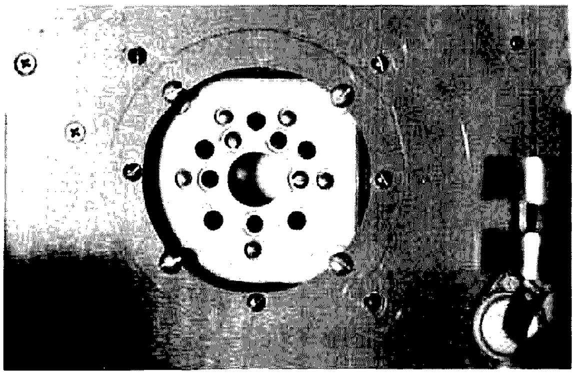
The amplifier design includes a dynamic bias circuit to bias the 3CX1200A7 tube beyond cut off between speech syllables on SSB or between dots and dashes on CW. This is especially useful when operating in a full break-in mode when the amplifier is placed in ready

Parts list for 3CX1200A7 amplifier (figs. 1-7).

item	description
B	*blower — Dayton 1C180*
cabinet	*10 × 17 × 14 inch, CTS model MCLS-101714, SPP-1014 side panels CP-1714 chassis panel (CTS IntraFab, 660 Lenfest Road., San Jose, California 65133)*
C1	*300 pF/10 kV vacuum variable capacitor*
C2	*1000 pF/3 kV vacuum variable capacitor*
C3	*1-8 pF miniature air variable capacitor*
CR	*HEP 170 diode or equivalent*
CR1	*9.1 volt, 1 watt zener diode (1N4739)*
CR2	*75 volt, 1 watt zener diode (1N4761)*
CR3	*100 volt, 1 watt zener diode (1N4764)*
CRB1,CRB3	*400 volt/4 ampere diode block (RS 276-1173)*
CRB2	*100 volt/4 ampere diode block (RS 276-1171)*
CRB4,CRB5	*diode strings, 10 HEP 170 diodes in series, each paralleled by a 470 kilohm resistor and a 0.01/1 kV disk capacitor*
L,C	*see table 2*
L1,L2,L3	*see table 1*
L4	*RF wattmeter pickup coil, 20 turns No. 22 enameled wire on T-50-2 ferrite toroid core*
M1	*1 ampere plate meter, Triplett 320-G*
M2	*100 μA meter, Triplett 320-G*
Q1	*MJ1000 NPN Darlington or equivalent*
Q2	*2N3055 NPN power transistor*
Q3	*2N3053 NPN transistor*
PC	*2 turns, 1/2 silver strap, 1-inch diameter 3 × 150-ohm 2-watt resistors*
RFC1	*90 turns No. 20 enameled wire on 3/4-inch form*
RFC2	*11 turns No. 14 enameled wire on 1/2-inch diameter air wound*
RFC3	*30 ampere bifilar filament choke, each coil is 16 turns on 1/2-inch ferrite rod*
RFC4,RFC8	*10 turns No. 14 enameled wire on 1/4-inch diameter ferrite rods*
RFC9	*1 mH/800 mA RF choke*
RL1,RL3	*4PDT Potter & Brumfield KHU17D11, 24VDC coil*
RL2	*SPDT vacuum relay, 26VDC coil*
RL4	*2PDT mercury plunger relay, Dayton 6X598-3*
RL5	*DPDT power relay, Potter & Brumfield PR-11-DY*
S1,S2	*2PDT push-button switches: ALCO 16TL5-22 ALCO 6T-4 yellow lens (S1) ALCO 6T-2 green lens (S2)*
S4	*momentary push-button switch, 1 pole/N.C. ALCO 16SL-11 switch ALCO 6S-2 red lens*
SCR	*2N1596 100 volt/1.6 ampere SCR*
T1	*filament transformer, 7.5 volt/21 ampere*
T2	*80 VAC (approx.) transformer, low current*
T3	*24 VAC/1 ampere transformer (Stancor P8661)*
T4	*1400 VAC-2 kVA power transformer*
TD1	*time delay relay, Amperite 115-N-180 (3 minute)*
Z1,Z2	*MOV transient protectors V130LA10A (RS 276-570)*

Notes:
1. Only the major items have been indicated in this parts list. See individual schematics to determine complete component complement.
2. The letter M indicated under the component value of capacitors indicates silver mica.





The 3CX1200A7 tube uses the same socket and chimney as the popular 3-500Z.

state at all times. The circuit can save up to 500 watts of power dissipation in the idle state.

Operation of the dynamic bias circuit is quite simple. RF is sampled through the 220 pF mica capacitor, C3. A voltage doubler is formed by the 1N914 diodes, providing a DC voltage to turn on the Darlington transistor Q1. Therefore, Q1 acts as a switch to return the bias voltage to the level determined by the zener diode CR1. The crowbar circuit formed by the SCR and zener diode CR2 is a protection circuit to prevent the amplifier from going into class C operation or the cathode voltage to rise toward the value of the plate voltage when enough drive power is applied should the Darlington transistor Q1 fail to conduct. A full theoretical treatment of this design is provided in reference 1.

Switch S3 provides a defeat of the dynamic bias circuit to set the operating bias level or defeat the circuit should the circuit fail in an open state. The circuit can be omitted if desired by replacing the bias module with a wire connecting the emitter of Q2 to the B-minus line (see **fig. 8**).

meter circuits

Metering is provided for plate current, grid current, high voltage and RF power output. Plate current is monitored at all times. This meter is in series with the B-minus lead. All other metered parameters are selected on a mulitmeter. The multimeter has a 100 mA

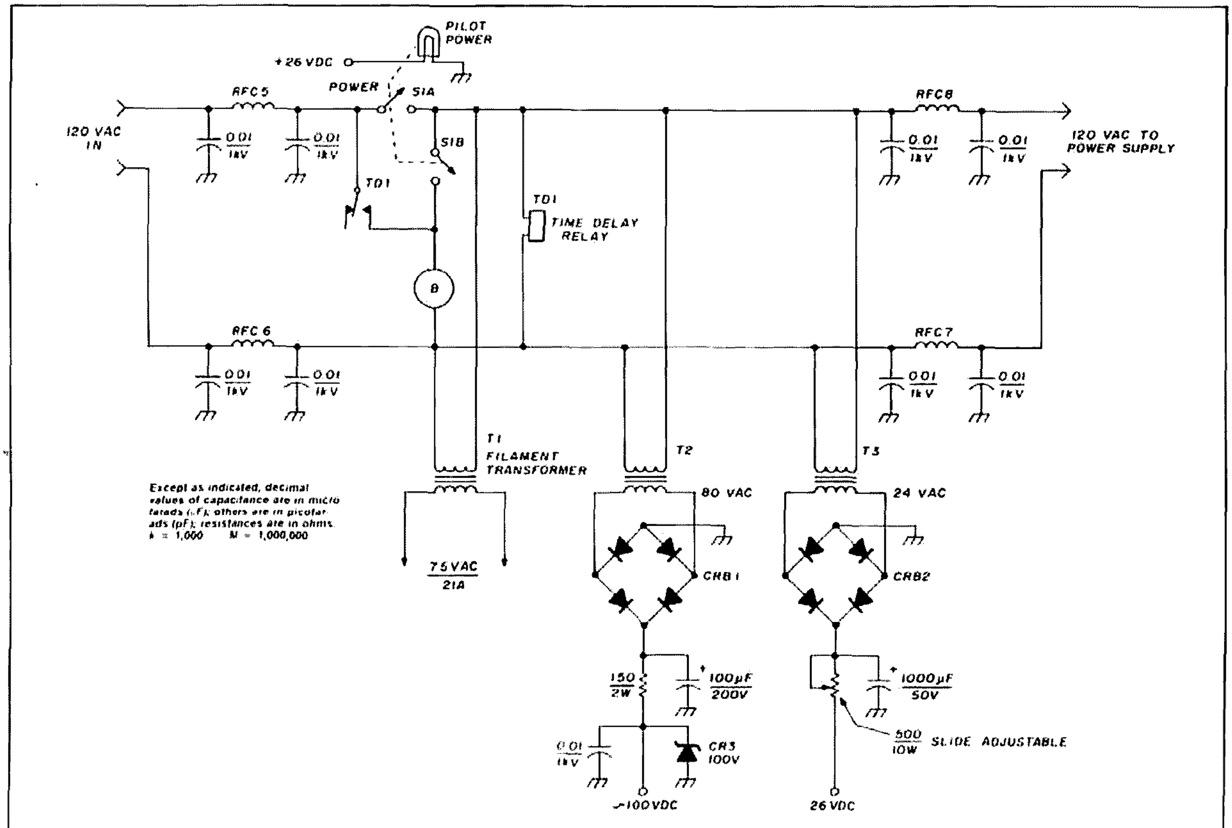
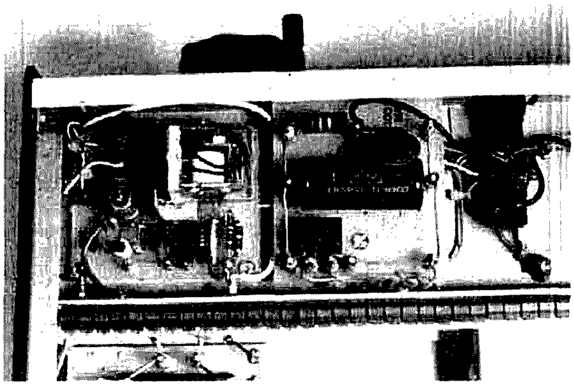


fig. 2. 3CX1200A7 control circuit.



Grid trip circuit (left side) and 100-volt bias supply (right side).

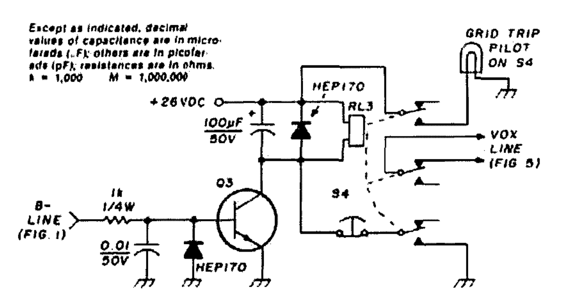


fig. 3. 3CX1200A7 grid trip circuit.

movement with proper calibration resistors for various scales. Any meter movement up to a 5 mA full scale is usable if an RF wattmeter is not desired. The wattmeter requires a very sensitive meter for proper operation.

Grid current is measured by monitoring the voltage across the 10-ohm resistor, R1, through which grid current is drawn. The 5-kilohm trim pot, R2, is used for proper calibration. R2 can be adjusted with a battery in series with a pot and current meter placed across R1 (see **fig. 4**) to vary and measure the amount of current drawn. R2 should be adjusted so that both meters read the same current.

High voltage is measured by monitoring a low voltage value developed by a voltage divider in the power supply. The sampled value is approximately 5 volts and coupled to the RF deck through the control cable.

Power is measured using a toroid-sensing RF wattmeter circuit. The wattmeter is shown in **fig. 5**. Three scales are provided: 200 watts forward, 2000 watts for-

ward, and 200 watts reflected. Voltage lines from the wattmeter into the front meter compartment use 0.001 feed-through capacitors to keep RF out. The wattmeter is in the RF line at all times even if the amplifier is not turned on.

Labeling the meters takes a lot of patience but really contributes to the appearance of the amplifier. It is necessary to choose a meter with an analog scale that has the correct number of divisions. However, the meter labeling makes no difference. In a very clean environment, remove the meter scale plate from the meter. Any markings on the meter can be removed with a pencil eraser. (Rub lightly but persistently.) The markings will come off the face plate, leaving a clean surface with an analog scale. The new number and letter markings can now be applied. I use dry transfer lettering to mark the meter scales to the desired values. The dry transfers now available from Radio Shack may be too large for small meters; varied assortments of smaller letters are available from most art supply stores.

RF relay sequencing

It is important to properly sequence the input RF relay, RL1, and the high power vacuum relay, RL2, to insure that the antenna is always connected to the amplifier before RF drive power is applied. This is accomplished by closing the output relay, RL2, slightly before the input relay RL1.

The timing circuit is shown in **fig. 6**. The 150-ohm resistor and the 100-μF capacitor form a time constant that delays closing of the input relay, RL1. The timing can be checked by applying a small voltage across the contacts of both RL1 and RL2 and watching the voltages on the scope as the amplifier is keyed up. The output relay, RL2, should close approximately 25 milliseconds before the input relay, RL1. The timing is adjusted by increasing the value of the 100-μF capacitor for longer delay or decreasing the value for shorter delay. The timing constants are totally dependent upon the relays used for RL1 and RL2. Avoid long time delays since a 50-ohm load isn't presented to the exciter when RL1 is open.

lead filtering

All control and power leads entering or leaving the RF deck are filtered with a pi-section circuit. The coils are made by winding ten turns of No. 16 enamel wire on 1/4 -inch diameter ferrite rods. Bypass capacitors are placed to ground at each end of the coil to form a pi-section. Locate each filter as close to the entry or exit point on the panel as possible. The line filters may not be necessary but provide insurance against RF running on leads outside the cabinet.

tank circuit

The tank circuit uses a pi-network design. The

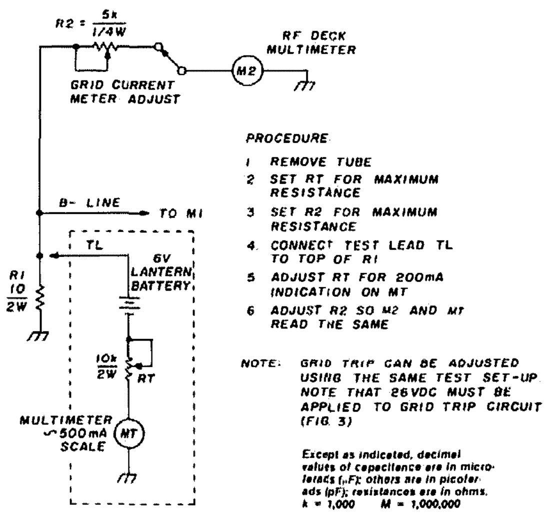


fig. 4. Set up for grid trip/grid current meter adjustments.

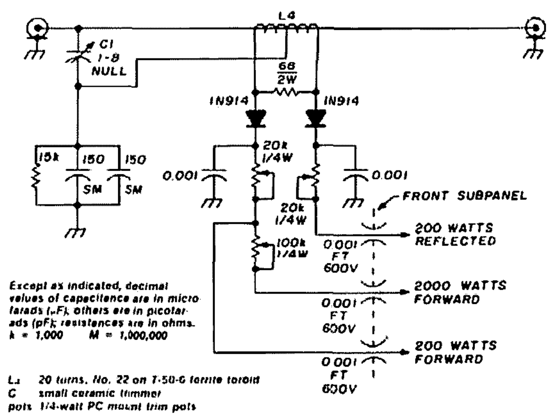
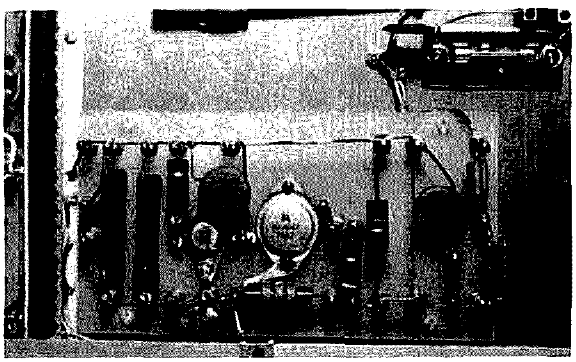


fig. 5. RF wattmeter circuit for the 3CX1200A7.



Dynamic bias module.

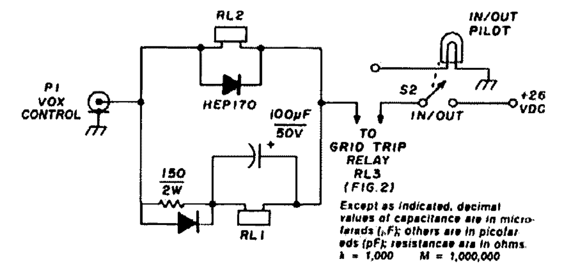
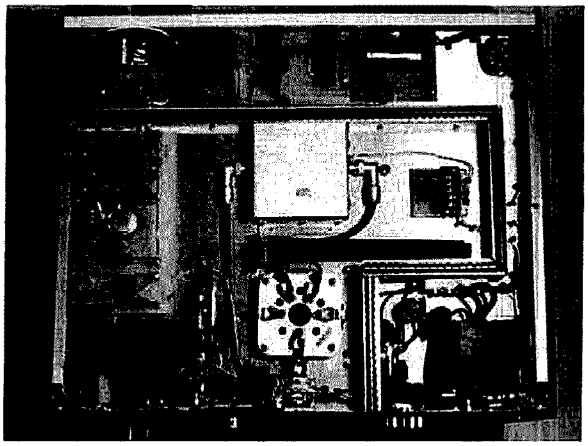


fig. 6. RF relay sequencing.



Bottom view of the amplifier illustrates the method used to isolate the RF circuitry from the control and power devices.

table 1. Tank circuit design parameters.

pi network:

F(MHz)	C1	C2	L	
				Q = 12
3.5	206	1174	9.95	R_L = 2500 ohms
7.0	107	608	5.15	
14.0	54	307	2.60	
21.0	36	204	1.74	
28.0	27	153	1.30	

coil: 4 turns, 1/4-inch tubing, 2-inch diameter
13 turns, 1/4-inch tubing, 3-1/4 inch diameter
taps: 28.0- 3 turns on 2-inch coil
21.0- 1 turn on 3-1/4-inch coil
14.0- 4 turns on 3-1/4-inch coil
7.0-11 turns on 3-1/4-inch coil
3.5 MHz toroid tank coil
3 × T200-2 cores taped
together with fiberglass tape
25 turns No. 12 enameled wire

pi-L network:

F(MHz)	C1	C2	L1	L2
3.5	244	1132	10.99	4.45
7.0	113	503	6.03	2.24
14.0	55	245	3.08	1.24
21.0	37	164	2.05	0.83
28.0	26	112	1.48	0.60

Q = 12
R_L = 2500 ohms

design was originally constructed with the 3-1000Z several years ago when the pi-network was the convention. Recent designs recommend a pi-L tank design because of a potential 15 dB improvement in attenuation of the second harmonic. However, the pi-L network needs a double pole band switch to select the inductance on both the pi-coil and the L-coil. Since a single pole switch was already in the amplifier and there was no evidence of TVI, the pi network was not retrofitted with a pi-L design. However, if this amplifier is built from "scratch," it is recommended that a pi-L design be used.

The design parameters for the pi and pi-L circuits are provided in **table 1**. The value for the plate impedance is 2500 ohms.

Vacuum variable capacitors are used for both the tune (C1) and load (C2) controls. Major advantages of vacuum capacitors include compactness, wide capacitance range with a low minimum capacitance necessary for 10 meters and tuning dial resettability. Equivalent air variables can be substituted. However, they are larger and their *Q* on 10 and 15 meters is lower.

input network

An input network is included to minimize distortion products and help insure that a 50-ohm load is presented to the exciter. The input network is a pi design with a Q = 1. **Table 2** summarizes the component values for each pi-section. The capacitors have been selected as standard values. The coils are wound on T50-2 toroid cores. The cores are large enough for the 100 watts drive power without overheating. However, if room permits, use T68-2 or even T75-2 cores to provide an extra safety margin.

The input network is built as a separate module (see input network pictorial) and ganged to the main band-

table 2. Input network design parameters

band	C1(pF)	L(µH)	C2(pF)	No. turns (T50-2)
80	860	2.15	860	23t/No. 20
40	440	1.11	440	17t/No. 20
20	220	0.56	220	12t/No. 18
15	150	0.38	150	9t/No. 14
10	110	0.28	110	8t/No. 14

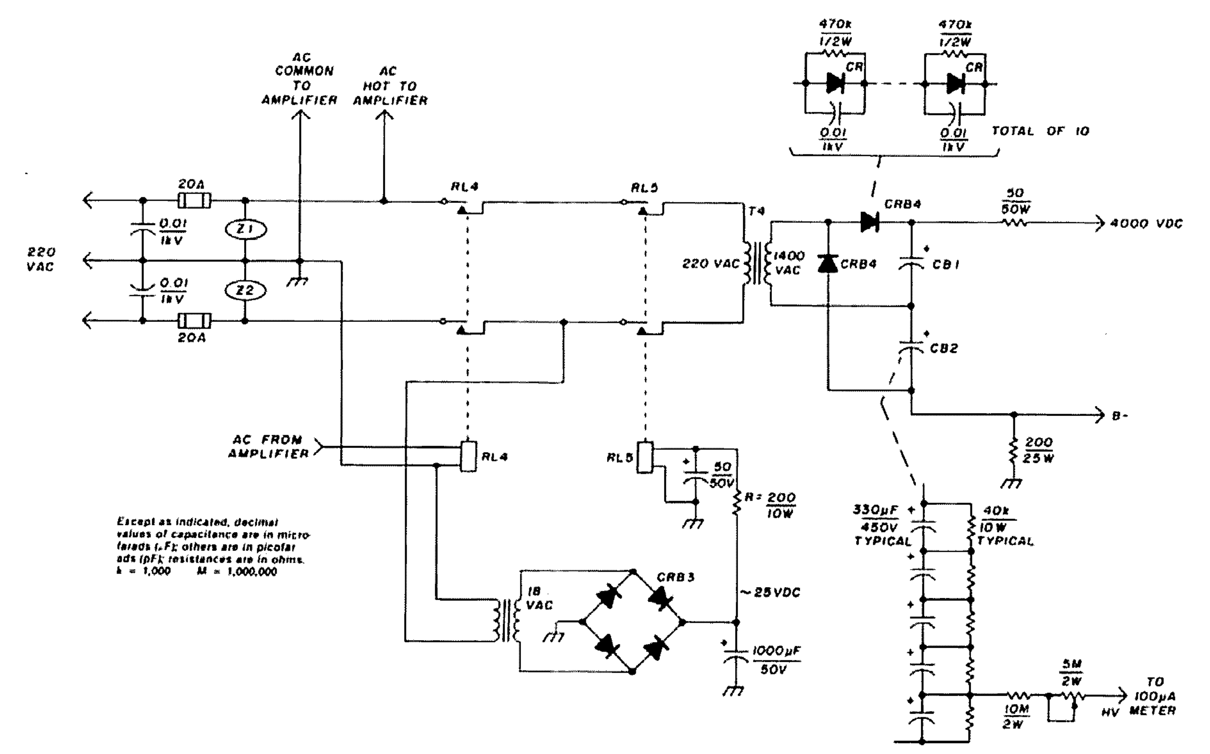
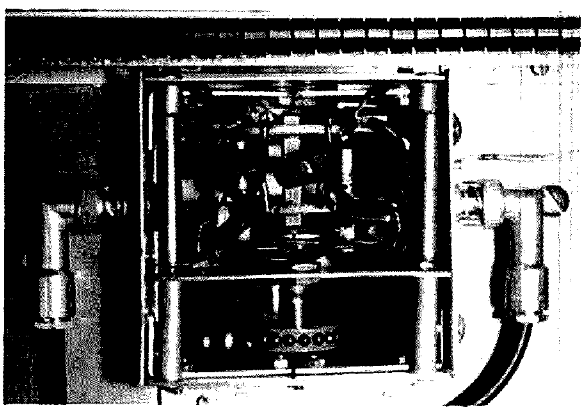


fig. 7. HV power supply.



Input network is built as a separate module.

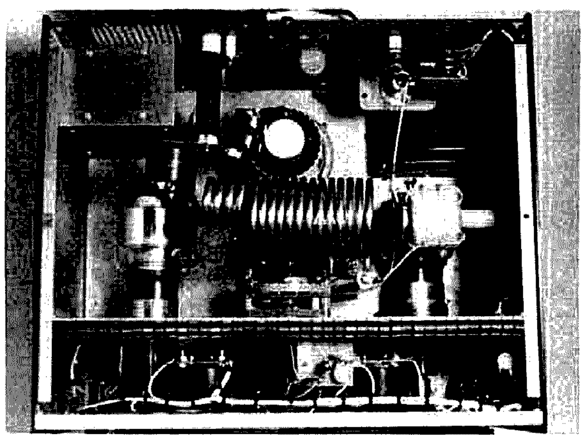
switch using a small chain sprocket available from reference 2. The unit was fully tested prior to installation into the amplifier.

cooling

Proper cooling is essential for long tube life. The 3CX1200A7 tube socket is mounted approximately 1/2 inch below the chassis. The under chassis is pressurized by a Dayton 1C180 blower mounted on the rear panel. The air flow is ducted by the tube chimney up around the base of the tube, through the external anode, and out the top of the amplifier. Refer to the 3CX1200A7 technical data sheet for more information on the cooling at different altitudes.[3]

HV power supply requirements

A good solid HV power supply is required to get the most out of any amplifier. This amplifier uses a 4000



Top view of amplifier illustrates small size of 3CX1200A7. Compare it to the vacuum variables in the lower section of the photo.

the ham notebook

a $100 printer for the Commodore 64

The interface and program for converting an ASR-33 teletype machine to an inexpensive printer for the VIC-20 ("VIC-20 Printer," ham notebook, September, 1984, page 88) can be adapted for use with the Commodore 64.

Figure 1A of the original article remains unchanged; in **fig. 1B**, only the labeling of the user's port is changed. **Figure 2** is replaced with a program listing that provides instructions to the Commodore 64.

Like its VIC-20 predecessor, the ASR-33 printer for the Commodore 64 produces typewriter-quality text appropriate for most data listing applications. Because ASR-33's can be found for as little as $50 to $75 — and because other parts and materials can be found in junk boxes or very cheaply — the total cost of this project should not exceed $100.

J.W. Dates, W2QLI

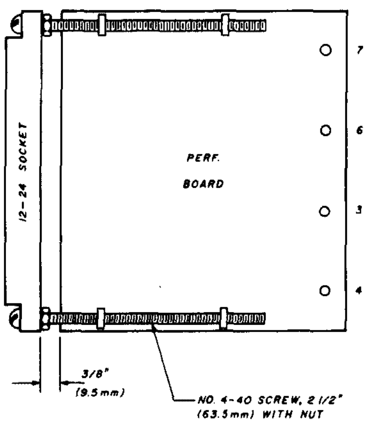
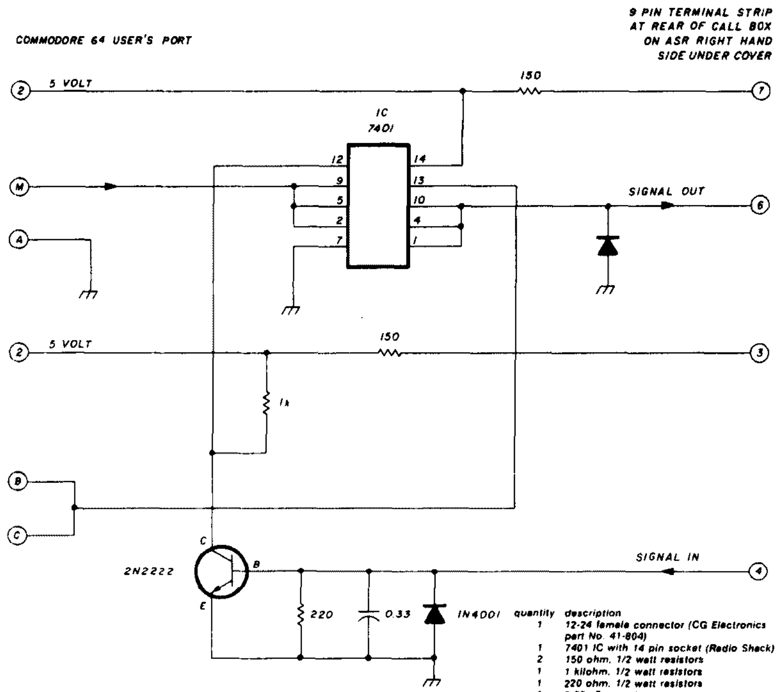


fig. 1A. Interface circuit uses multi-pin edge connector to match C-64 requirements.



quantity	description
1	12-24 female connector (CG Electronics part No. 41-804)
1	7401 IC with 14 pin socket (Radio Shack)
2	150 ohm, 1/2 watt resistors
1	1 kilohm, 1/2 watt resistors
1	220 ohm, 1/2 watt resistors
1	0.33 μF capacitor
2	1N4001 diode
1	2N2222 transistor
1	piece of perforated board 2 × 2-1/2 inches 0.100 × 0.100 hole spacing
2	4-40 × 2-1/2 inch screws with nuts

fig. 1B. Interface circuit that ties a C-64 to an ASR-33.

```
10 REM ASR 33 TTY
15 REM FILE#>128 FORCR WITH LF
20 REM 163=2STOP,7 BIT ASCII,110 BAUD
100 OPEN129,2,3,CHR$(163)+CHR$(224)
110 GET#129,A$
200 REM MAIN LOOP
210 GET B$
220 IF B$<>""THEN IF B$=CHR$(13)THEN
     PRINT#129,B$;CHR$(0);CHR$(0);:GOTO230
225 IF B$<>""THENPRINT#129,B$;
230 GET#129,C$:IFC$<>""THEN PRINT#129,C$;
240 PRINT B$;C$;
250 SR=ST: IF SR=0 OR SR=8 THEN200
300 REM ERROR REPORTING
310 PRINT "ERROR";
320 IFSR AND 1 THEN PRINT "PARITY"
330 IFSR AND 2 THEN PRINT "FRAME"
340 IF SR AND4 THEN PRINT"REC BUFF FULL
    "350 IF SR AND8   THEN PRINT "BREAK"
360 IF(PEEK(673)AND 1)=0 THEN360
370 CLOSE129:END

READY.
```

fig. 2. C-64/ASR-33 program listing.

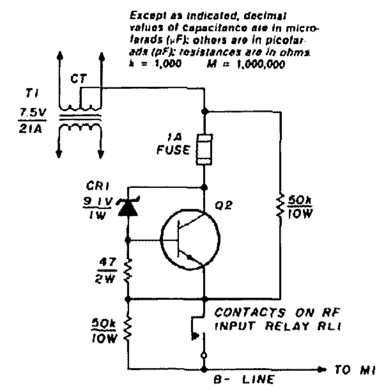
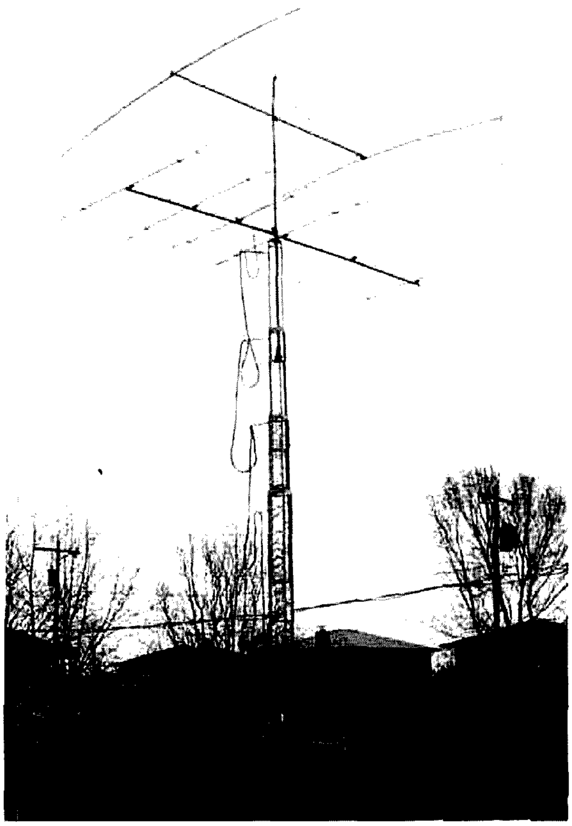


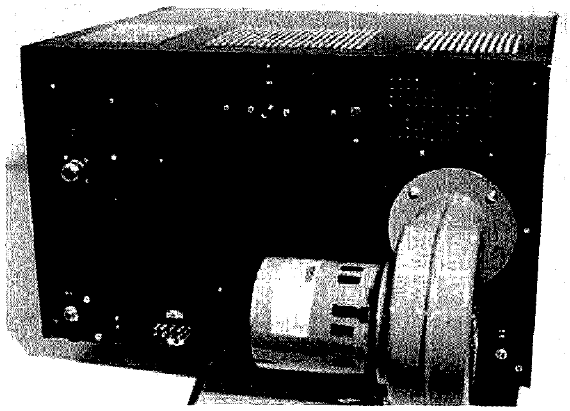
fig. 8. Alternate circuit for conventional bias control.

volt power supply. The HV drops to 3600 volts when 800 mA is drawn.

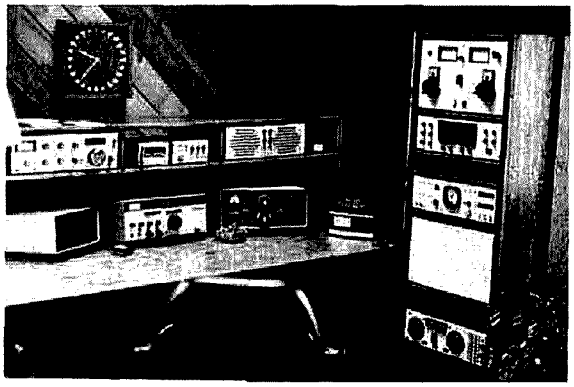
The power supply is built as a separate unit that sits behind the operating desk. Its circuit is shown in fig. 7.



Set in 7½ cubic yards of concrete, author's 70-foot Tri-ex 470D tower is completely self-supporting, with fully motorized crank-up, tilt-over capability. Lower antenna is a KLM KT34XA; upper antenna, at 80 feet, is a two-element Cushcraft 40-meter beam. Both were erected by K8RA without assistance.



Amplifier rear view. Notice the dynamic bias defeat switch in the lower right corner.



K8RA's station is largely home-brewed. Shelf holds receiver described in August, 1983, edition of *QST*; construction details will appear in W6SAI's forthcoming revision of *The Radio Handbook*. To the right are two homebrewed keyers and a speaker to match the amplifier. On desk top are a second speaker, a Collins KWM 380 transceiver, a Collins 30L1 linear amplifier, a rotor control and a 2-meter rig. Rack, at right, contains homebrewed equipment exclusively; from top to bottom, the 3CX1200A7 amplifier described in this article, a station monitoring oscilloscope, a station control console, and an antenna tuner with built-in memory. Blank panel marks space for new 8877 linear amplifier currently under construction. (No, Jerry didn't build the clock; it was a $25 "find" at Dayton. But he did build the 16 × 22-foot solid cedar addition to house his station. Skylights, not pictured here, afford overhead view of antennas.)

amplifier tuning

Amplifier adjustment is initally done into a 50-ohm dummy load. With filament and operating plate voltage applied, grid and plate current meters should read zero when the amplifier is not keyed up by the exciter. Shorting the VOX in/out line engages the antenna relays which switches the amplifier into the RF line. Static plate current should still be zero if the dynamic bias circuit is used. If the dynamic bias circuit is not

included or bypassed with the defeat switch, S3, the static plate current should rest at approximately 150 mA. If grid current is observed with the plate or tune capacitors in any position, it is probably a sign of parasitic oscillations and should be corrected.[4]

Drive power can now be slowly increased until the plate current is approximately 250 mA. The Tune and Load controls are adjusted for maximum power output as indicated on the RF power output meter. Excitation is now increased and the plate and load tuning capacitors adjusted until approximately 800 mA plate current with 200 mA grid current is achieved, with maximum output indicated. When the above conditions have been met, the loading should be increased slightly to insure proper linearity. The RF power output should not exceed 1500 watts. Use a scope to monitor signal quality if available. Under SSB speech conditions, peak plate current with no clipping or compression will kick to about 400 mA and grid current to about 100 mA.

conclusion

The amplifier has been in operation for over six months and has operated flawlessly. As expected, another fine tube by EIMAC.

references

1. J.A. Bryant, W4UX, "Electronic Bias Switching for RF Power Amplifiers," *QST*, May, 1974, page 36.
2. *Precision Mechanical Components*, Winfred M. Berg, Inc., 499 Ocean Avenue, East Rockaway, New York 11518.
3. *3CX1200A7 Technical Data Sheet*, Varian/EIMAC, 301 Industrial Way, San Carlos, California 94070.
4. William Orr, W6SAI, *Radio Handbook*, 20th Edition, Howard W. Sams, Inc., Indianapolis, Indiana. (See "Low-frequency Parasitic Suppression," page 17.20.) (*The 22nd Edition of* Radio Handbook *is available from Ham Radio's Bookstore, Greenville, New Hampshire 03048, for $12.95 plus $3.50 shipping and handling.*)

ham radio

ham radio TECHNIQUES

Bill Orr, W6SAI

big signals from the north

Have you heard those outstanding signals from Finland? Every active DXer has heard these block-busters in DX contests. Why? Because our friends up north have no hesitation about erecting *giant* antennas.

Consider the antenna system that Marti, OH2BH, uses. A photo of his "Christmas Tree" systems and rotator are shown in **figs. 1** and **2**. The 140-foot (42 meter) rotatable steel tower was manufactured and installed by OH8QD. It is guyed by nonmetallic Parafil ropes which terminate on slip rings. A heavy duty ART-8000 rotator supports the system.

A brace of six-element KLM beams is mounted on the tower — six over six on 20 meters at the 140-foot (42.6 meter) and 170-foot (51.8 meter) levels. Single six-element beams for 21 and 28 MHz are at the 120-foot (36.5 meter) and 95-foot (29-meter) levels. Antenna switching is accomplished by a set of remotely controlled vacuum relays located at the 105-foot (32-meter) level.

The relay allows a selection of either one or both of the antennas. Three SPST (single pole, single throw) relays are used in **fig. 3**. The relays are shown in the "off" position. Coax line L1 is an electrical 1/4-wavelength of 75-ohm coax and transforms the nominal 50-ohm antenna feedpoint resistance to about 112 ohms. The two 112-ohm ports at relay 2 are put in parallel and become a 56-ohm termination that closely matches the 50-ohm system design value. The relays are in a weatherproof box mounted halfway between the antennas. Fifty-ohm lines from the box run to each antenna.

A representation of the relay wiring in the box is shown in **fig. 4**, as is the switching table located at the operating position.

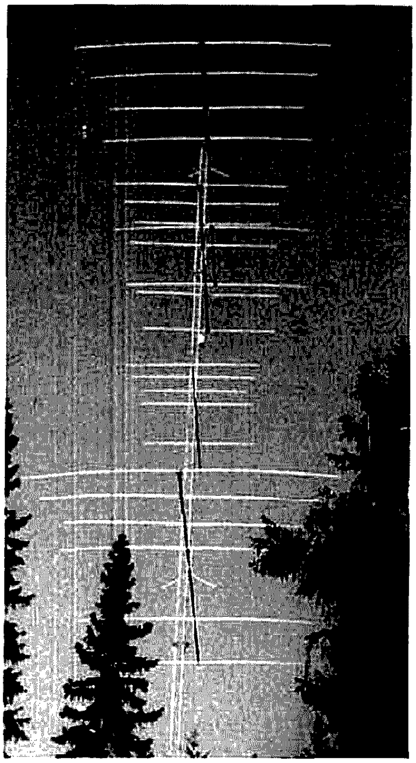


fig. 1. The "Christmas tree" antenna at the contest site of Marty, OH2BH. Six-over-six elements on 20 meters, backed up by single six-element beams on 21 and 28 MHz are mounted on the 170-foot high tower. The 20-meter beams can be switched in combinations from the operating position.

To reduce vibration in the antenna system, which could be destructive in harsh winter storms, each element has a 5-foot (1.5 meter) length of rope cemented into the inside of the outer element tips.

Another interesting antenna installation is that of Simon, OH8OS. A close-up view is shown in **fig. 5** and the slip-ring assembly is shown in **fig. 6**. The antenna is composed of eight six-element beams, stacked two-over-two! OH6JW is believed to have a similar setup. The tower sits on a rotator and the guys are attached to slip rings at appropriate levels.

No doubt other big arrays exist. I'd like to receive pictures of them to put in this column!

how about 160 meters?

Even though there aren't any stacked, rotary arrays on 160 meters that I know of, there are certainly some big signals from well-known DXers on this band. One of the prominent signals on this band (and on others) comes from Jay, AD8C. Jay has a 106-foot (32.3 meter) Rohn tower, guyed at three levels with 1/4-inch (0.6 cm) diameter stranded steel wire. Only the top set of guys is broken by insulators at 25-foot (7.6 meter) intervals. A length of heavy-wall aluminum tubing projects 15 feet (4.5 meters) above the top of the tower. The tower is turned by a WØMLY-modified prop-pitch rotator.

On 160 meters, the tower is shunt-fed with an 82-foot (25 meter) length of electrical conduit as the gamma rod, spaced 1.5 feet (0.45 meter) from the

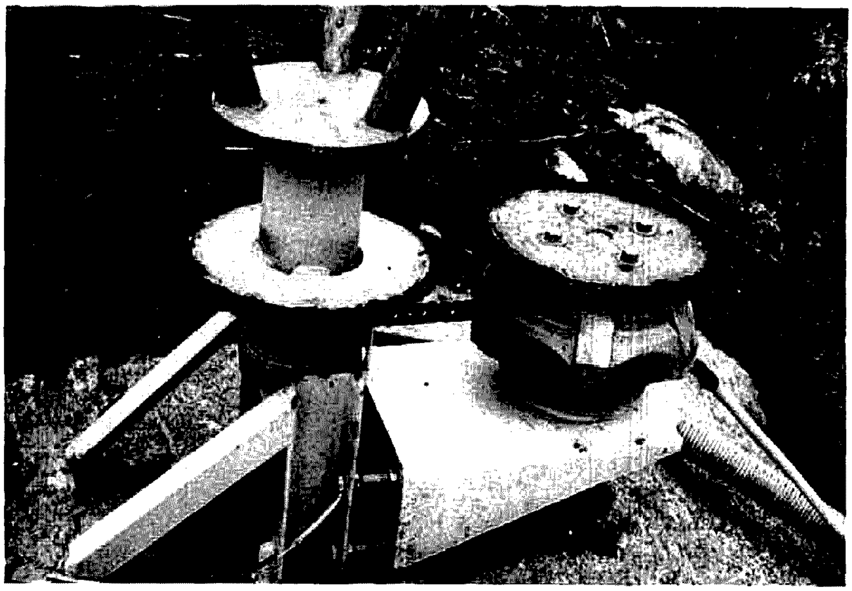
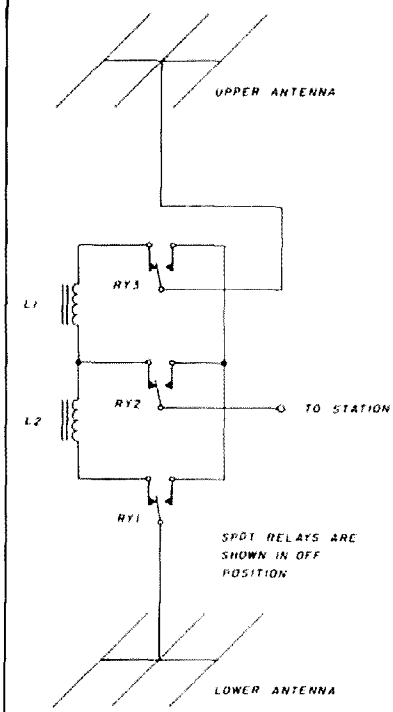


fig. 2. The 170-foot high tower of OH2BH is rotatable. This is view of the base support and the offset rotating mechanism. The tower is guyed by nonmetallic Parafil ropes which terminate on slip rings.



relay	both	upper	lower
1	off	off	on
2	off	on	on
3	off	on	off

fig. 3. Antenna switching system at OH2BH permits selection of lower, upper, or both antennas.

tower. Thirty radials are used, with 15 of them a full quarter-wavelength long. A 5kV, 1000 pF vacuum variable capacitor is used as the gamma capacitor. The 2-to-1 bandwidth of the antenna is about 55 kHz.

As for the other bands, Jay uses five phased slopers on 80 meters, patterned after the 40-meter system described in the ARRL *Antenna Handbook* (1974 edition). It provides about 3 dB forward gain and a front-to-back ratio of about 10 to 15 dB.

The rest of the antennas include: a three-element Yagi for 7 MHz at 107 feet (32.6 meters), a four-element Yagi for 14 MHz at 120 feet (36.5 meters), and four-element Yagis for 15 and 10 meters at 55 and 113 feet (16.7 and 34.4 meters). To complete the "antenna farm," there is an eight-element Yagi for 2 meters at 95 feet (29 meters).

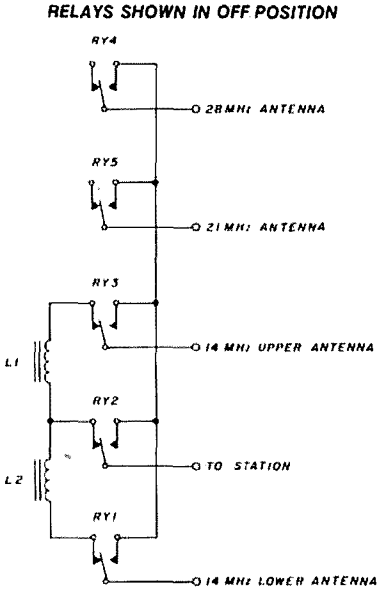
Jay says he is satisfied with the installation that it has withstood severe winter storms, and does a "credible job" in DX pileups, particularly on the lower bands.

One of the outstanding 160-meter signals from Asia comes from Kuni, JA7NI, who lives in a three-story apartment house. Atop the building Kuni has a 46-foot (14 meter) high heavy wall aluminum mast (**fig. 7**). The mast is top-loaded by a 76-foot (23 meter) length of wire that slopes downward to a short pole mounted on an adjacent building. The aluminum roof trim of his building is used as a ground connection. The antenna works well; Amateurs in all US districts have been contacted, with many QSOs made along the east coast, including WA2SPL, W1FC, W2FJ, K2EK, KC2SB, W3ESU, and others.

After getting this information, I wrote Joe, WA2SPL, to see what antennas he's used for DX work to Japan. He told me that his first 160-meter antenna was an inverted-vee with the apex at 150 feet (45.7 meters) and the ends at 75 feet (22.8 meters). Joe says that although it was a "pretty good" antenna for transmitting, precipitation static during snow storms made the antenna useless for receiving. He then put up a 1000-foot (348 meter) Beverage antenna for receiving that worked very well. On the common assumption that "if big is good, bigger is better," Joe erected a 1500-foot (457 meter) Beverage wire that proved to be even better than the shorter one.

The next step was to raise the inverted-vee to 200 feet (61 meters), with the ends at 90 feet (27.4 meters). This produced an immediate improvement in his signal — to the point at which he couldn't hear some of the DX that called him! So Joe went back to modifying the Beverage receiving antenna.

It was obvious that more and better Beverages were needed, so Joe put up a 2100-foot (640 meter) wire for Europe, a 1500-foot (457 meter) wire aimed at Japan, a 1000-foot (348 meter) wire for South America, a 1,000-foot wire on the Caribbean and a 3100-foot (944 meter) wire aimed at Australia/New Zealand. (Where do these guys get the space to put up these antennas?)



relay	14 both	14 upper	14 lower	21	28
1	off	off	on	off	off
2	off	on	on	on	on
3	off	on	off	off	off
4	off	off	off	off	on
5	off	off	off	on	off

fig. 4. Complete schematic of OH2BH antenna selection relays. L1 and L2 are 1/4-wavelength sections of 75-ohm coxial cable.

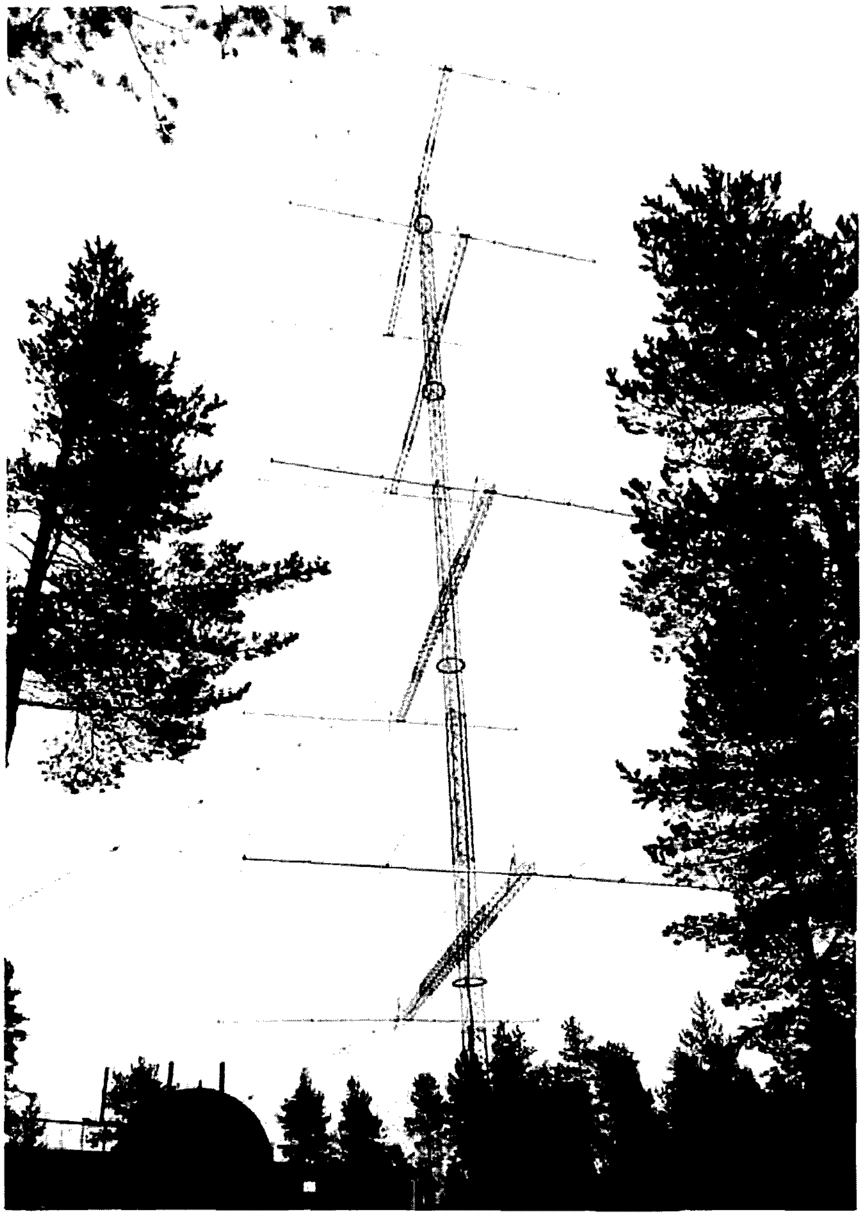


fig. 5. The monster 20-meter array of Simon, OH8OS. The antenna consists of eight six-element beams, stacked two-over-two. Simon puts through an S-9+ signal on the West Coast of the USA when other Europeans are inaudible. The complete tower is rotatable.

After a few months Joe found out that a Beverage antenna could be too long. He pruned the 3100-foot wire back to 1500 feet (457 meters) and it worked much better.

With each antenna, a 20 dB preamplifier was used to bring signals up to good copy. By this time, Joe had 150 countries to his credit on 160 meters!

Even with his success with the Vee and the Beverages, Joe was curious about the luck his DX friends had with vertical antennas on 160 meters. For various reasons, he couldn't shunt-feed his 200-foot tower, so he dropped a rope down on a 45-degree angle to the ground and lowered a wire to ground level from the 130 foot (40 meter) elevation on the rope. He placed 24 quarter-wave radials beneath the vertical antenna.

He soon discovered, by direct comparison, that under no conditions did the vertical perform better than the high Vee antenna. He was sure that the vertical was relatively worthless until he later met Dana, W1CF, and went to Dana's station to work it during a *CQ* Worldwide DX contest. The station was equipped with two 1200-foot (365 meter) Beverage receiving antennas, plus two Col•Atch•Co 61-foot (18.6 meter) verticals with 80 radials under each one. The contest results were amazing — the W1CF station ended up with 830 QSOs and a score of 323,000 points!

Now Joe is wondering why his full-size vertical seemed to perform so poorly. Too few radials? Low conductivity ground in the vicinity of the antenna? He may soon have the answer. He's moving to a new location, and by the time you read this, he may have put up his new 160-meter antenna system — four full-size, phased

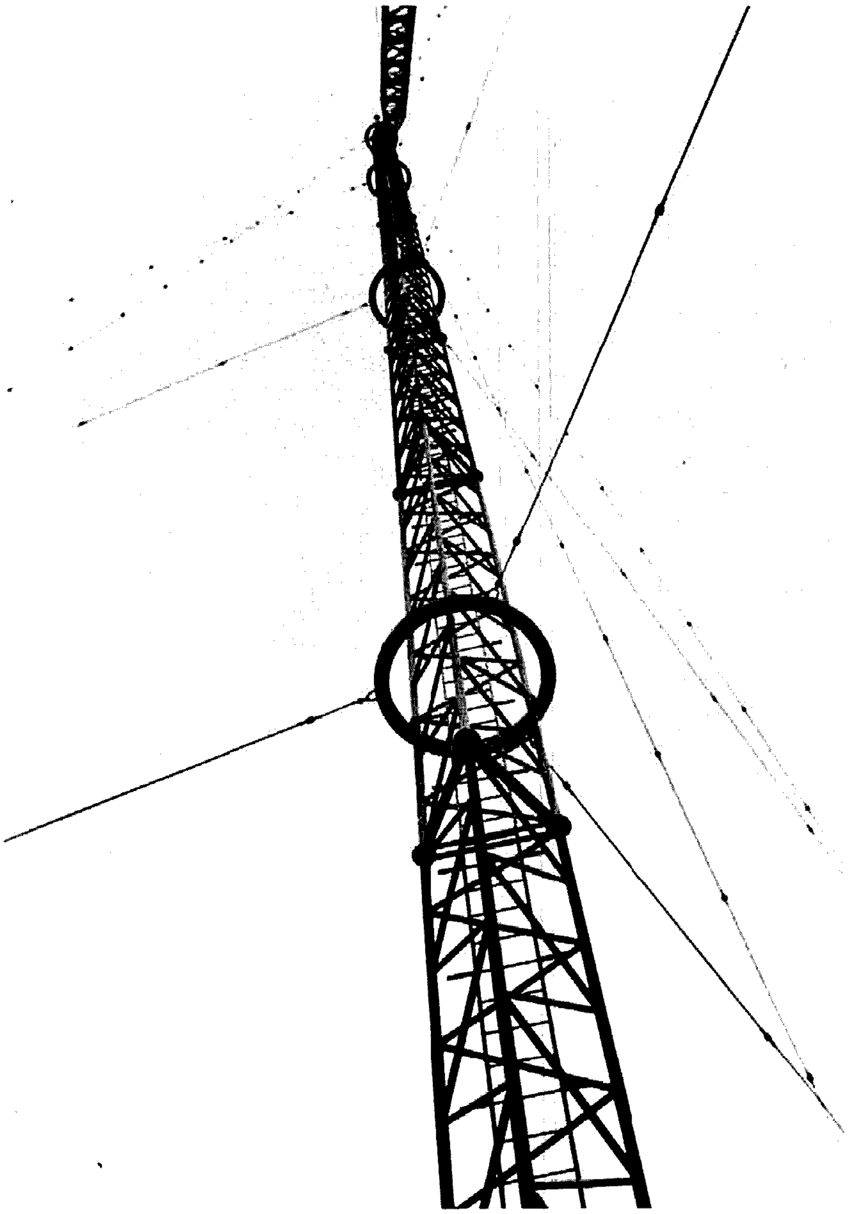


fig. 6. Slip-ring installation on tower of OH8OS. Would you care to climb the ladder to the top array? (Visible at top of picture is the arm of the crane ready to lift the antennas to their final positions on the tower.)

verticals on a mountain in northern Vermont. JA7NI, keep your ears open for this one!

You *can* be a topnotch 160-meter DXer with only a simple antenna. Bob, VE1YX, says, "No super-antenna here! I have a simple, coax-fed dipole, 60 feet high on a hilltop about 450 feet above sea level. So far, 135 countries — the best DX being VK6HD long path at 2100Z."

long-wave hams in Australia

Sixteen years ago, before WARC 79, the United States considered an Amateur band in the 160-190 kHz region. The idea had merit, but opposition arose because this was the range used by carrier current data transmissions on power lines and also by long-wave European broadcasting stations. Sadly, the idea was dropped before WARC 79 started.

Even so, John (VK3ACA) and Pet (VK3QI) pursued the idea with th Department of Communications i Australia and, after a pause, they wer issued experimental licenses for lov frequency operation on 196 kHz (153 meters). John received the call sig AX3T35 and Peter became AX3T36. third Amateur, Dennis (VK3WV), joi ed them and got the call sign VL3 (**fig. 8**). The stations all ran about 10 watts input.

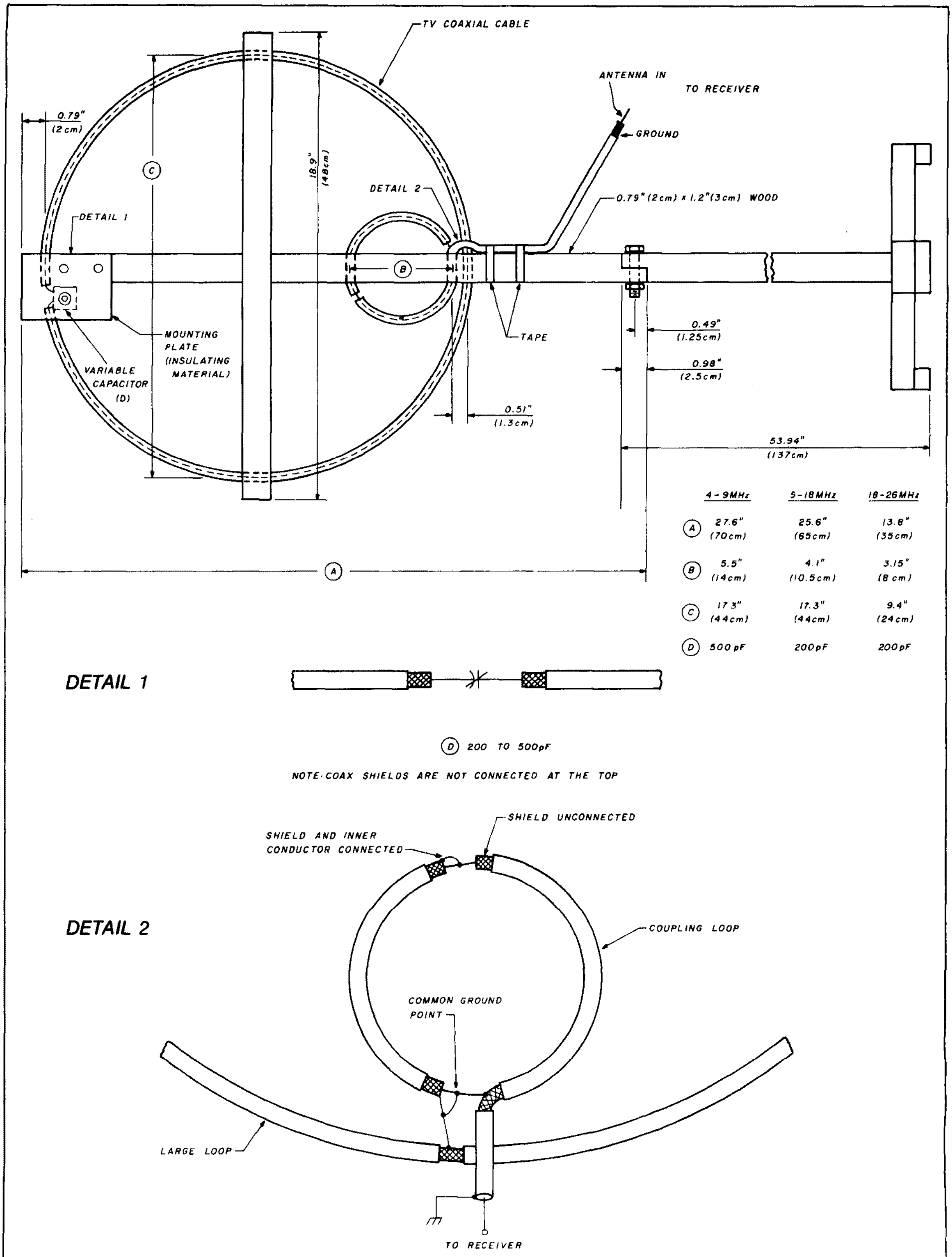
Antennas were a problem. AX3T3 used a 30-foot vertical with a hug base loading coil and an extensiv grounding system. He estimated h antenna efficiency was about 0.37 pe cent. Station VL3Y used a 135-fo wire, which wasn't much better. Eve with the poor antennas, the first co tact was made in April, 1981. Co tinued contacts between the three se tions showed that the propagatio range via ground wave was about 30 miles during daylight, and possib more during the hours of darkness.

AX3T35 is writing an article, sched uled to appear shortly in the journal o the Wireless Institute of Australia *Amateur Radio*, about the experience of the only three licensed low-fre quency ham staions in the worl (Good work, lads — I'll put a shrim on the barbie for you!)

an anti-jamming HF loop antenna

People other than Radio Amateu are interested in efficient and effectiv HF antennas. Radio Free Europe, i particular, tries in every way to com bat the Soviet jamming that plague their broadcasts. In this regard, the have published information on buildin a simple and inexpensive direction receiving antenna that can be helpf in reducing jamming interferenc Details of this antenna are shown i (**fig. 9**.)

The antenna consists of a shielde loop made of coax cable (RG-58/U, o RG-59/U, for example). A smaller loo couples the tuned loop to the receive To hold the cables in the loop form a wooden support in the shape of cross is used. The coax is passe



	4-9MHz	9-18MHz	18-26MHz
A	27.6" (70cm)	25.6" (65cm)	13.8" (35cm)
B	5.5" (14cm)	4.1" (10.5cm)	3.15" (8 cm)
C	17.3" (44cm)	17.3" (44cm)	9.4" (24cm)
D	500pF	200pF	200pF

fig. 9. Receiving loop recommended by "Radio Free Europe." (Dimensions provided in centimeters.)

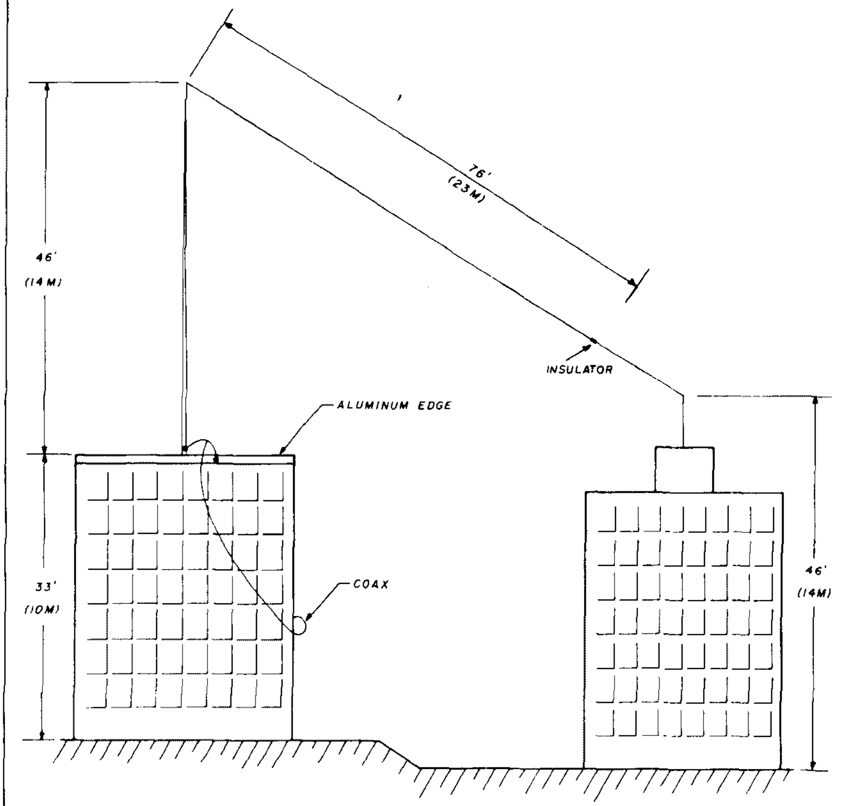


fig. 7. Apartment antenna of 160-meter DX operator Kuni, JA7NI, consists of top-loaded 46-foot (14 meter) mast atop a three-story apartment building. Shield of coax is grounded to aluminum roof trim.

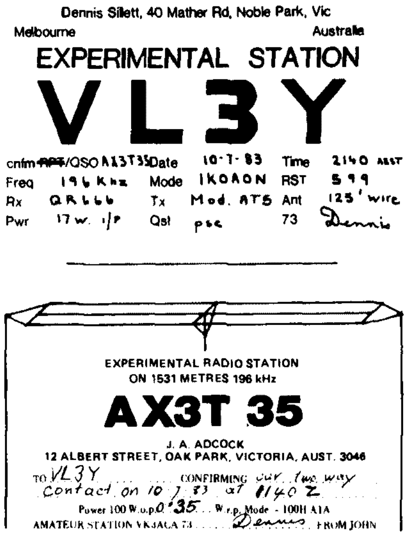


fig. 8. QSL cards of low-frequency experimental Amateur stations in Australia.

through appropriate holes in the structure. A variable capacitor at the open end of the loop tunes it to the required frequency. The coupling link matches the symmetrical, balanced loop to the unbalanced coax line to the receiver.

The vertical portion of the support structure is hinged to allow tilting the antenna for improvement of the rejection null. The antenna can be rotated horizontally to provide both azimuth and elevation alignment.

To cover the shortwave spectrum from 4 to 26 MHz, three loop sizes are required, as shown in **fig. 9**.

To use the loop antenna, receiver and antenna are tuned for maximum signal at the desired frequency. The antenna is then rotated and tilted to minimize the interfering signal. A reduction of 30-to-1 in jamming strength is predicted for short wave listeners, depending upon the location of the jamming station with respect to the desired station.

While Radio Amateurs may not use the loop for reduction of intentional jamming, this design may prove to be of benefit for operation in DX contests where it's helpful to null out loud local competition.

Because of the reduced pickup of the loop compared to a full-size antenna, it may be necessary to add a preamplifier between loop and receiver to bring signals up to full strength.

the new *Beam Antenna Handbook*

For the past two years I've been absent from the bands, devoting every moment of my spare time to completing, with Stu, W2LX, the extensive revision of the *Beam Antenna Handbook*. It was a lengthy job; the old text was ripped apart and new text prepared. New illustrations were added. New, up-to-date antenna dimensions, based upon recent computer studies conducted on Yagi arrays were cataloged. The result, after much hard work, is a completely new edition of *Beam.**

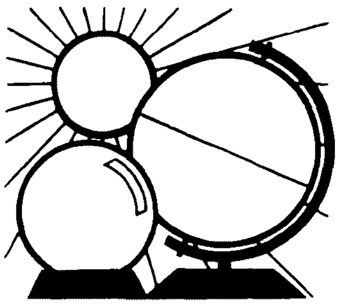
The book includes new data on HF Yagi antennas (two to five elements), element spacing, and the effect of element taper. There's new information on erecting beams and general installation data as well. VHF long Yagis are covered, together with complete design tables for the home constructor. Complete English and metric dimensions are given for all antennas.

There's helpful information on feed systems and SWR measurements. A systematic test procedure is provided to help you determine whether your beam is operating properly. The information on checking the accuracy of your SWR meter is worth the price of admission!

Now that the book is ready, perhaps I'll have time to get on the air! Or will another project come along?

*Available from Ham Radio's Bookstore, Greenville, New Hampshire 03048, $9.95 plus $2.50 shipping and handling.

ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

signal levels during the summer months

Absorption results in loss of energy from the signal as it collides with ions on its path through the D region, 60-80 miles (100-120 km) above the earth. How much energy is absorbed per transit of the D region depends on the location of the sun, and is a function of cosine X, the zenith angle to the sun. Maximum absorption occurs at the subsolar point, directly under the sun; absorption decreases as the signal path moves away from the subsolar point in any direction.

On any propagation path, absorption increases with the number of crossings of the D region and also varies inversely with frequency. Therefore, in working DX it pays to use the higher frequency bands and obtain consequently greater distances per hop.

In a study of several midlatitude communication links of varied lengths over the years of this 10.7 solar cycle, some general trends were noticed. The ionosphere, a balanced energy system tends to return to a "normal" level of ion density after each new solar perturbation. It is a fact that between sunspot numbers 20 and 120, signal absorption changes only by approximately 8 dB for a one-hop path 2500 km in length. The one-hop path is the easiest in which to see changes. When the signal travels three or more hops, the changes get blurred between the hops. There is more absorption with the larger number of hops, but the effect of absorption per hop is not linearly additive. Each additional hop subtracts less energy from the signal.

Absorption is inversely proportional to frequency when the angle the signal wave makes with the ionospheric layer is constant, as in vertical ionospheric sounding. However, this frequency dependency is hard to assess because as the frequency changes so does the extent of layer penetration, thereby changing the incidence angle somewhat. As a rough estimate, 10 MHz signals tend to incur twice as much signal absorption as 20 MHz signals. However the largest absorption effect occurs between night and day; it is so great it tends to mask out the measurement of the secondary causes. The midlatitude communication link paths showed 10-30 dB of signal loss between night and day in most seasons, with paths near the equator surpassing even these. These different absorption effects add up to give an overall signal loss of 120 dB on the average.

What can DXers do to enhance their effectiveness during summertime operations? Review the chart on the next page for the highest band available to the DX area you wish to contact. Operate on towards evening, taking advantage of its lower absorption, but before the maximum usable frequency drops off very much. Then use the graph provided in the February, 1985, column that shows take-off angle (TOA) vs. ionosphere height. Use the height chart provided, too. The lowest TOA means fewer hops and less absorption of your signal. Make sure your antenna radiates substantial energy at that low TOA in order to give you the best chance of contacting the desired DX station.

Know the current conditions (in terms of signal absorption and variability, QSB) by listening to radio station WWV on 5, 10, 15 MHz at 18 minutes after the hour. If the solar flux has just increased, absorption will be high. In addition, potential fading conditions (QSB) are associated with an A figure of greater than 15 or a K figure greater than 4. These indicate pronounced signal absorption on the higher latitude paths. These clues can help you be your own forecaster — therefore a better DX operator.

last-minute forecast

During August noise will be up and signal strength down, however sporadic-E propagation will occur. This just means that we can expect good DX conditions on the higher frequency bands in the third and last weeks of the month while the lower bands are expected to be best the first week of the month. It's always helpful to check WWV broadcasts to confirm these conditions. For the VHF/UHF enthusiast the moon's perigee will occur on the 19th, with a full moon on the 30th. The Perseids meteor shower will occur from the 10th to 14th, with a maximum rate expected on the 11th and 12th, with better than fifty meteors per hour. This is an excellent shower. Meteor shower activity has already been reported as heavier than usual this year.

band-by-band summary

Six-meter paths will open for a half hour to a couple of hours on some days around local noon. Sporadic-E propagation will make this short-skip path possible out to nearly 1200 miles (2000 km) per hop.

Ten and fifteen meters will have a few short-skip E_s openings and some long-skip openings to southern areas of the world during daylight, though only during periods of high solar flux. Some transequatorial (TE) openings associated with mildly disturbed geomagnetic-ionospheric conditions may occur in the evening hours toward the end of the month.

Twenty, thirty, and forty meters will have DX from most areas of the world

AUGUST

WESTERN USA

GMT	PDT	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)
0000	5:00	20	20	15	15	20	15	15*	20
0100	6:00	20	20	20	15	20	15	15	20
0200	7:00	20	20	20	15	20	15	15	20
0300	8:00	20	20	20	15	20	15	15	20
0400	9:00	20	20	20	20	20	15*	15	20
0500	10:00	20	20	20	20	30	15*	15	20
0600	11:00	20	_30_	20	_20_	30	15*	15	20
0700	12:00	_20_	_30_	_20_	_20_	30	15	15	20
0800	1:00	_20_	_30_	_20_	_20_	_30_	15	20	20
0900	2:00	_30_	_30_	_30_*	_20_	_30_	15	20	20
1000	3:00	_30_	_30_	_20_	_30_	_30_	20	20	30
1100	4:00	_30_	_30_	_20_	_20_	_30_	20	20	_30_
1200	5:00	_30_	20	20	20	_30_	_20_	_20_	_30_
1300	6:00	_30_	20	15	20	_30_	_20_	_20_	_30_
1400	7:00	_30_	20	15	20	20	_20_	_20_	_30_
1500	8:00	_30_	20	15	15	20	_20_	_20_	_30_
1600	9:00	_30_	20	15	15	20	_30_	_20_	_30_
1700	10:00	20	20	15	15	20	_30_	_30_	_30_
1800	11:00	20	20	15	15	20	_30_	_30_	30
1900	12:00	20	20	15	15	20	20	20	20
2000	1:00	20	20	15	15	20	15	20	20
2100	2:00	20	20	15	15	20	15	15	20
2200	3:00	20	20	15	15	20	15	15	20
2300	4:00	20	20	15	15	20	15	15*	20

MID USA

GMT	MDT	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)	CDT
0000	6:00	20	20	20	15	20	15*	15	20	7:00
0100	7:00	15	20	20	15	20	15	15	20	8:00
0200	8:00	15	20	20	15	20	15	15	20	9:00
0300	9:00	20	20	20	20	20	15	15*	20	10:00
0400	10:00	20	20	20	20	20	15	15	20	11:00
0500	11:00	20	20	_30_	20	30	15	15	20	12:00
0600	12:00	_30_	_30_	_30_	_20_	30	20	15	20	1:00
0700	1:00	_30_	_30_	_20_	_20_	_30_	20	20	20	2:00
0800	2:00	_30_	_30_	_30_	_20_	_30_	20	20	30	3:00
0900	3:00	_30_	_30_	_20_	_20_	_30_	20	20	_30_	4:00
1000	4:00	_30_	_30_	_20_	_30_	_30_	_20_	20	_30_	5:00
1100	5:00	_30_	_30_	20	_20_	_30_	_20_	_20_	_30_	6:00
1200	6:00	_30_	20	15	20	_30_	_30_	_20_	_30_	7:00
1300	7:00	_30_	20	15	20	_30_	_20_	_20_	_30_	8:00
1400	8:00	20	20	15	20	20	_20_	_20_	_30_	9:00
1500	9:00	20	20	15	15	20	_30_	_20_	_30_	10:00
1600	10:00	20	20	15	15	20	_20_	_20_	_30_	11:00
1700	11:00	20	20	15	20	20	15	_30_	20	12:00
1800	12:00	20	20	15	20	20	15	_30_	20	1:00
1900	1:00	20	20	20	15	20	15	20	20	2:00
2000	2:00	20	20	20	15	20	15	20	20	3:00
2100	3:00	20	20	20	15	20	15	20	20	4:00
2200	4:00	20	20	20	15*	20	15	15	20	5:00
2300	5:00	20	20	20	15	20	15*	15	20	6:00

EASTERN USA

GMT	EDT	N (Asia, Far East)	NE (Europe)	E (S. Africa)	SE (Caribbean, S. America)	S (Antarctica)	SW (New Zealand)	W (Oceania, Australia)	NW (Japan)
0000	8:00	20	20	20	15	20	15*	15	20
0100	9:00	30	20	20	15	20	15	15	20
0200	10:00	_30_	20	20	20	20	15	15	20
0300	11:00	_30_	20	20	20	20	15	15	20
0400	12:00	_30_	20	_30_	20	30	20	15	20
0500	1:00	_30_	20	_30_	_20_	30	20	15	20
0600	2:00	_30_	_30_	_20_	_20_	30	20	15	30
0700	3:00	_30_	_30_	_20_	_20_	_30_	20	20	_30_
0800	4:00	_30_	_30_	20	_20_	_30_	_20_	20	_30_
0900	5:00	30	_30_	20	_30_	_30_	_20_	20	_30_
1000	6:00	20	_30_	15	_30_	_30_	_20_	20	_30_
1100	7:00	20	_30_	15	_20_	_30_	_30_	_20_	_30_
1200	8:00	20	_30_	15	20	_30_	_20_	_20_	_30_
1300	9:00	20	20	15	20	_30_	_20_	_20_	_30_
1400	10:00	20	20	15	15	20	20	_20_	_30_
1500	11:00	20	20	15	15	20	20	_30_	20
1600	12:00	20	20	15	15	20	15	_30_	20
1700	1:00	20	20	15	15	20	15	_20_	20
1800	2:00	20	20	15	15	20	15	20	20
1900	3:00	20	20	15	15	20	15	20	20
2000	4:00	20	20	15	15	20	15	15	20
2100	5:00	20	20	20	15	20	15	15	20
2200	6:00	20	20	20	15	20	15	15	20
2300	7:00	20	20	20	15	20	15*	15	20

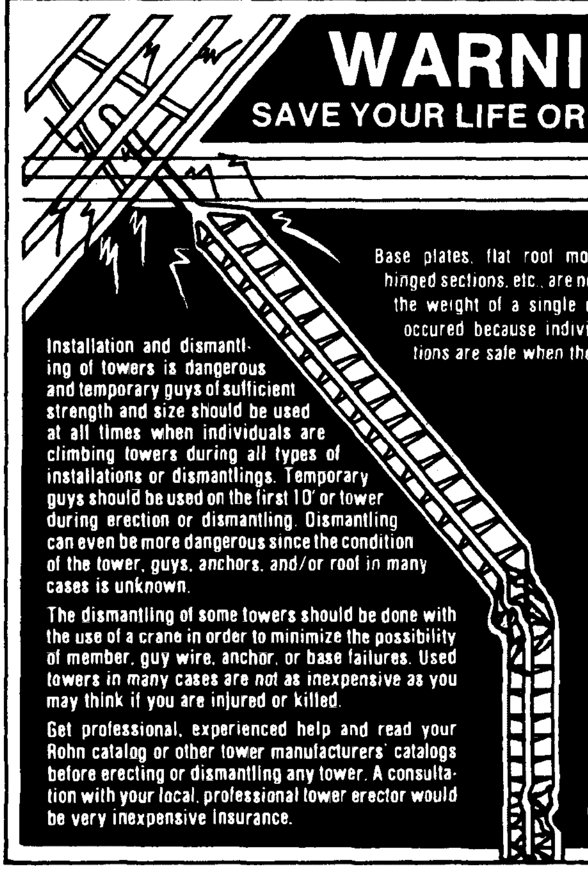
The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

during the daytime and into the evening hours almost every day. Forty meters will join this daytime group toward the end of the month because of lower signal absorption (lower lowest usable frequency) during these sunspot minimum years. During unusually high 27-day solar flux days, 40 meters may not be usable; both 40 and 30 meters may not be usable at the predawn hours after the high flux day. The DX on these bands may be either long-skip to 2500 miles (4000 km) or short skip E_s to 1250 miles (2000 km) per hop. The length of daylight will be decreasing noticeably at the end of the month.

Thirty, forty, eighty, and one-sixty are all good for nighttime DX, even though the background thunderstorm noise may be severe, especially on 80 and 160 meters. Propagation of signals via sporadic-E layer may occur and assist around sunset. This will be the final month of the E_s season.

ham radio



NEW products

900-MHz scanner converter

Hamtronics, Inc. has announced a new converter for scanner radios to cover the 900-MHz land mobile band. An adaptation of the very popular CVR-806, which covers the 806-896 MHz band, the new 900-MHz converter allows coverage of new services now being assigned or proposed for the 880-960 MHz range, including additional land mobile services, such as police and fire departments, government and non-government fixed stations, industrial, scientific, and medical services, and the proposed 902-928 MHz Amateur band. Also included are proposed new cellular telephone and paging services and existing and new broadcast studio-transmitter links. The price of the CVR-900 is $88 plus $3 for shipping and handling. Other converters are available for the 72-76, 135-144, 240-270, 400-420, and 806-896 MHz bands at the same price.

For a complete catalog of Hamtronics' products, send $1 to Hamtronics, Inc., 65-F Moul Road, Hilton, New York, 14468-9535. (For overseas mailing, please send $2 or 4 IRCs.)

UHF cavity amplifiers

Varian/EIMAC has introduced six new UHF cavity power amplifiers designed for FM, CW, pulse, or single-sideband linear service in the 280 to 530 MHz frequency range.

Using the EIMAC 3CX800A7 high-μ power triode, the cavity amplifiers eliminate equipment design complications and extra power supplies associated with UHF tetrode cavities, yet provide comparable stage gain. Power gain in FM or CW service for all cavities is on the order of 11 dB with efficiency ratings in excess of 55 percent.

In addition to being more efficient, the new cavity amplifiers offer reliability in the targeted bands because of the comparatively simple design.

The cavities provide approximately 450 watts power output in CW and FM service over the following ranges: CV-2401, 390 to 450 MHz; CV-2402, 375 to 420 MHz; CV-2403, 280 to 300 MHz; CV-2404, 470 to 530 MHz; CV-2405, 330 to 370 MHz; and CV-2406, 450 to 470 MHz.

Standard 50-ohm Type N input and output RF connectors are used for all cavities. Silver-plated components are used to ensure the best performance and efficiency.

The cavities are forced-air cooled and designed for mounting to a customer's 19-inch panel. Each has a net weight of about 13 pounds. All are 14 inches wide, 10 inches deep, and range in height from 6.2 to 9.3 inches.

For additional information or literature, contact Varian/EIMAC, 301 Industrial Way, San Carlos, California 94070.

Circle #308 on Reader Service Card.

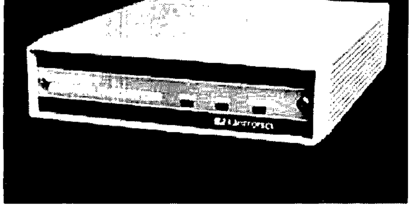
packet communicator

Kantronics has announced a new product for Amateurs using computers in the shack: the Kantronics Packet Communicator.

Interest in packet radio has grown in recent years with volunteer Amateur groups doing research and testing of the new mode. With the ARRL adoption of the AX.25 protocol as the Amateur standard, packet radio became a viable form of data exchange. Thousands of Amateurs have proven the new mode reliable using a hardware and software program devised by the Tuscon Area Packet Radio group (TAPR).

To better utilize the new packet technology Kantronics has designed a new hardware format for processing the packet protocol. By using an internal microprocessor to handle the protocol, and integrated circuits for signal processing, the Kantronics Packet Communicator becomes the most compact and inexpensive finished packet unit available today.

Data is transmitted between the Kantronics Packet Communicator and the computer using a Series RS232 or TTL port. Baud rates of 300, 1200, and 9600 can be used. Any terminal or communications software program can be used to set up the computer to communicate with the Packet Communicator. Special Packet Terminal (PAC-Term™) programs for many popular personal computers will be available soon from Kantronics.



System compatability, the ability to exchange data with existing Packet Terminal Node Controllers, has been achieved with the Kantronics Packet Communicator by using the popular TAPR software.

Almost all of the commands and operation procedures used by the TAPR group are used with the Kantronics Packet Communicator. Both the ARRL standard AX.25 and Vancouver protocols are incorporated in the unit. The Kantronics Packet Communicator supports baud rates of 300, 400, 600, and 1200, but the unit does not support full duplex operation.

An added feature of the Kantronics Packet Communicator is the ability to select either Bell 103 or 202 tones for 300 baud operation. This will allow the operator to switch to the lower tone set, improving performance at slower speeds on the HF bands. This feature makes the Kantronics Packet Communicator an excellent choice for gateway use on the HF bands. The suggested retail price is $389.95.

For further information, contact your local Kantronics dealer, or Kantronics, 1202 East 23rd Street, Lawrence, Kansas 66046.

Circle #307 on Reader Service Card.

headset/boom microphone for TR-720

Communications Specialists has announced the availability of a high quality headset/boom microphone that plugs directly into the TR-720 handheld airband transceiver. The CS-65 HEAD-

SET/MIC was developed to permit improved transmission and reception with the TR-720 in noisy environments. This new accessory is light (12 ounces) and features cushioned, noise-attenuating ear pads that can be adjusted for a comfortable fit. A flexible boom supports and electret noise-cancelling microphone. Supplied with a 5-foot cable to connect to the radio, the unit comes with a push-to-talk switch attached. The CS-65 HEADSET/MIC lists for $69.95.

For further information, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665.

Circle #306 on Reader Service Card.

NEW products

touchtone decoder kit

Engineering Consulting has announced the model TSD 4-digit sequential touchtone (DTMF) decoder for mobile and base station use. Each board can have a unique 4-digit user-programmable access code. The board will install in most VHF/UHF and HF gear, allowing alarm or mute functions to be implemented. Typical applications include muting the audio circuit until a valid 4-digit code is received. An alarm can be sounded upon receipt of the access code to alert the operator that someone is calling on the channel, without having to listen for your call sign.

The model TSD decoder is easy to install in any 12-volt radio. Speaker audio or low-level audio may be used to listen to the tones. An open collector transistor provides the output control to switch a small relay or alarm device. Either momentary or latching control is provided with jumper selection. Upon receipt of the access code a latch or momentary pulse is provided. Send the code again and the latch will reset. The TSD is available wired and tested from Engineering Consulting for $59.95.

For further information, contact Engineering Consulting, 583 Candlewood Street, Brea, California 92621.

Circle #305 on Reader Service Card.

super CW

Super-CW, a CW audio processor from Hildreth Engineering, includes an 8th order Butterworth cascade of staggered pairs to provide excellent skirt rejection without excessive response to impulse noise. The 3 dB passband is from 700 to 800 Hz with a 3 to 30 dB shape factor of less than 3.

More than just a filter, the unit features an S/N BOOST function, which is driven by the pre-filter, to provide a signal-to-noise ratio enhancement of over 10 dB as compared with the linear filter position. It does this for signals that are well below the noise in a typical 3 kHz audio bandwidth. This boost circuitry uses compound-complex filter/limiter/filter elements with added active circuits (patent applied for) that creates S/N enhancement for CW – or any pulse-code-modulation (PCM) signal – analogous with that enjoyed by FM communications systems. A second and very important benefit is ear protection. When in the S/N BOOST position, the sudden onset of strong signals or noise pulses just can't happen. You get a clean, distortion-free signal at a sound pressure level uniquely determined by the AF GAIN control.

A 2-watt power amplifier with a controlled voltage gain of 25 is included to allow a reduction in receiver RF gain, which reduces the tendency toward non-linear disturbances in your receiver's IF and/or product detector when listening to a weak signal under the condition of strong signal QRM outside of the 700 to 800 Hz passband. The unit receives its input from your receiver's speaker output. Power supply requirements are 12 to 15 VDC at a nominal 350 mA peak. The unit will drive a 4 to 8 ohm speaker.

For more information, contact Hildreth Engineering Corporation, P.O. Box 60003, Sunnyvale, California 94088.

Circle #303 on Reader Service Card.

filters for TS-940S

Matched sets of filters for the Kenwood TS-940S are available from International Radio, Inc. The SSB-2.1 kHz set consists of one 8.83 MHz, 2.1 kHz drop-in, 8-pole crystal filter and one 2.1 kHz 455 kHz 8-pole crystal filter (wired in). This matched set will provide an overall system selectivity of 2.0 kHz at 6 dB and 2.5 kHz at 60 dB. The shape factor is 1.25.

The CW-400 Hz matched set consists of one drop-in 8.8 MHz 400 Hz 8-pole crystal filter and one drop-in 455 kHz filter, for system selectivity of 400 Hz at 6 dB and 700 Hz or less at 60 dB. The shape factor is 1.75 or less.

Sets are priced at $139.00 each, or $260 for both. All crystal filters are guaranteed for two years. Quantity discounts are available.

For further information, contact International Radio Inc., 364 Kilpatrick Avenue, Port St. Lucie, Florida 33452.

Circle #304 on Reader Service Card.

solar breakthrough

The ENCON Corporation has introduced the first commercially available amorphous (thin film) photovoltaic to the Amateur Radio marketplace. Genesis,™ a state-of-the-art 5-watt PV module, represents a breakthrough in solar cell technology. Typical applications for the new modules, which may be wired together for increased power, include battery maintenance on ham equipment, recreational vehicles, and boats. The modules can produce enough power for telecommunication from QRP stations, security equipment, and some home lighting.

Manufactured by ARCO Solar, the Genesis is designed for easy use. Convenient front mount-

NEW products

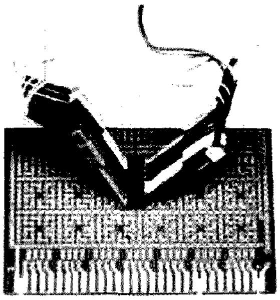
ing holes and a long lead wire make the Genesis module easy to install and use. Modules bolt directly to almost any secure surface and produce power even in low or diffused light. Power specifications for the Genesis include: voltage at maximum power, 14.5 VDC; current at maximum power, 0.35 amperes; open-circuit voltage, 20.8 VDC; and short-circuit current, 0.435 amperes. Carrying a one-year warranty and retailing for under $100, its advanced technology provides an affordable solar alternative accessible to everyone.

For more information, contact ENCON Corporation, 27600 Schoolcraft, Livonia, Michigan 48150.

Circle #302 on Reader Service Card.

pico hook adapter with wire/cable plug-on capability

Designed specifically for narrow test contact access, the E-Z HOOK P-25 "pico Hook" adapter features an ultra-thin blade and housing to allow connector stacking and ease of use, while allowing the flexibility of independent wire and cable assembly.



Incorporated into the design is a one-piece, gold-plated beryllium copper conductor and hook for assured signal accuracy, qualified to MIL-G-45204. The P-25 permits connection to 0.025-inch square or 0.030-inch round push-on jumpers and is designed for connection directly to a single leg of a DIP or other components which have a maximum diameter of 0.025 inch.

Lightweight and Finger-eze Hypo Action permit hook-up to delicate wires where weight and leverage may otherwise damage components. The barrier on the tip prevents shorting between DIP legs. A durable plastic body and plunger, and stainless steel compression spring provide approximately eight ounces of contact pressure.

For details, contact E-Z hook, Inc., P. O. Box 450, Arcadia, California 91006-0450.

Circle #301 on Reader Service Card.

flea market

RATES Noncommercial ads 10¢ per word; commercial ads 60¢ per word **both payable in advance**. No cash discounts or agency commissions allowed.

HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing) on a space available basis only. Repeat insertions of hamfest ads pay the non-commercial rate.

COPY No special layout or arrangements available. Material should be typewritten or clearly printed (**not** all capitals) and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** cannot check each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue.

DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

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Herts EN6 3JW
England

Holland Radio
143 Greenway
Greenside, Johannesburg
Republic of South Africa

Coming Events

ACTIVITIES

"Places to go..."

KENTUCKY: The Central Kentucky ARRL Hamfest, sponsored by the Bluegrass ARS, Sunday, August 11, 8 AM to 5 PM, Scott County HS, Longlick Road and US 25, Georgetown. Tech forums, license exams, awards and exhibits. Air conditioned facilities. Free outdoor flea market space. Tickets $3.50 advance and $4.00 at gate. Talk in on 76/16. For information or tickets SASE to Scott Hackney, KI4LE, 629 Craig Lane, Georgetown, KY 40324.

PENNSYLVANIA: The Tioga County Amateur Radio Club's 9th annual Hamfest, Sunday, August 18, Island Park, Blossburg. 9 AM to 5 PM. VE's will give walk-in exams. For information write TCARC, PO Box 56, Mansfield, PA 16933. Flea market, dealers, park and pool, snack bar. Admission $3.00. Spouse and kids free. Talk in on 146-19/79, 146-52/52 and CB. For information: Durwood Learn, WB3DKZ, 11 Bryden Street, Wellsboro, PA 16901. (717) 724-5613.

TEXAS: The Austin ARC and the Austin Repeater Association in conjunction with the Texas VHF-FM Society announces the third annual Austin Summerfest, August 2, 3 and 4, Austin Marriott Hotel, I-35 and US 290, Austin. Seminars, QCWA hospitality suite, dealer displays. FCC exams for all classes Saturday. Ladie's programs. Austin's "Aquafest" is this same weekend and will provide additional entertainment. Registration $5 advance, $7 at the door. Persons under 18 admitted free. Swapfest tables available starting 6 AM Saturday at $1 per table, limit two. Talk in on WA5YAN/R 146.34/94. For more information: Austin Summerfest, PO Box 13473, Austin, TX 78711.

PENNSYLANIA: The 48th annual South Hills Brasspounders and Modulators Hamfest, August 4, 9 AM to 4 PM, South Campus, Community College of Allegheny County, Pittsburgh. Tickets $3 each. 2/$5. Oscar, RTTY and packet radio forums. Flea market. Talk in on 146.13/73 and 146.52 simplex. For more information: Bill Gardiner, 4756 Child Drive, Pittsburgh, PA 15236.

INDIANA: The annual Bloomington Hamfest, Sunday, September 1, 147.18/.78 repeater site, Vernal Pike off SR 37 bypass. 8 AM to 2 PM. Admission $2.00. Food concession. No charge for setups, bring your own tables. For more information: Bob Myers, K9KTH, 306 S. Fairview Street, Bloomington, IN 47401. SASE or dall (812) 332-1105.

INDIANA: The 6th annual Grant County Amateur Radio Club Hamfest, Sunday, August 11, 4H Fairgrounds, Marion. Doors open 8 AM. Refreshments, free parking, license exams. Table reservations $2/8' table. Donation $2.00 advance. $3.00 gate. For more information SASE to WB9EAP, Brooks Clark, 2202 South Boots Street, Marion, IN 46953.

WASHINGTON: Tacoma Hamfair, sponsored by the Radio Club of Tacoma. August 17 and 18, Pacific Lutheran University, Tacoma. Tech seminars, forums, travelogs and more. Large flea market. License exams, send 610 to W7BUN. Registration $5.00. Dinner $8.00. Flea market table $15/day; $20/2 days includes one registration. Register with Grace Tietzel, AD7S, PO Box 45079, Tacoma, WA 98445 or call Eva Anderson, WB7QNS, (206) 564-8347.

OHIO: 43rd annual Findlay Hamfest sponsored by the Findlay Radio Club at the Hancock County Fairgounds. Sunday, September 8, 6:30 AM to 5 PM. Advance tickets $3.00 by September 1. At the door $4.00. Tables $6.00 each. Outdoor flea market spaces $3.00 each. Talk in on 147.75/15. For more information write Findlay Radio Club, PO Box 587, Findlay, Ohio 45839.

MISSOURI: The St. Charles Amateur Radio Club's Hamfest '85, August 25, St. Charles City Hall complex, 200 North 2nd Street, St. Charles. Giant flea market, commercial vendors, XYL programs, FCC exams, food available. All under cover. Parking $1.00, tickets $1.00 advance; $1.50 door. Talk in on 146.07/67 and 146.52. Tickets from WD0CZE, 121 Barkwood Trail, St. Charles, MO 63303.

CALIFORNIA: Valley of the Moon Amateur Radio Club's fifth annual "Ham" breakfast and swapmeet, Sunday, August 11, Sonoma Community Center, 276 East Napa Street, Sonoma. 9 AM to 4 PM. Breakfast 9 to 11:30 AM. Sausage, eggs, pancakes, 'taters, o.j. and coffee, all you can eat for $5.00! Swap tables setup starts 8 AM, spaces $5.00 each. Better bring your own tables. Open auction at 1 PM. Surrounding points of interest for the whole family. Admission $1.00. For reservations or more information: Darrel Jones, WD6BOR, 358 Patten St., Sonoma, CA 95476. (707) 996-4494.

VERMONT: The annual BARC International Hamfest, August 10 and 11, Old Lantern Campgrounds, Charlotte. $4.00 both days. Children under 12 free. Outdoor flea market space $2.00. Indoors $5.00. RC model airplane show. CAN-AM tug-o-war. Talk in on 34/94, 01/61 and 52. Queries to Roger, WA1OZE; flea market info Bob, W1DQO. Both at Box 312, Burlington, VT 05402.

ILLINOIS: The Shawnee Amateur Radio Association is sponsoring SARA Hamfest '85 Sunday, September 8, John A. Logan College Gym, Highway 13 near Carterville. New equipment and computers, ladies' activities, displays, flea market, crafts. All inside. FCC exams Sunday AM. Lunch available. Admission $3.00. Talk in on 146.25/.85, 146.52 simplex, 3.925 MHz. For information: Shawnee ARA, 502 West Kenicott, Carbondale, IL 62901. (618) 457-7586.

1985 BLOSSOMLAND BLAST, Sunday, October 6, 1985. Write "BLAST", PO Box 175, St. Joseph, MI 49085.

MAINE: The 1985 Windsor Hamfest, Saturday, September 7, Windsor Fairgrounds. Flea market, programs, speakers, distributors, and the traditional Saturday bean and casserole supper. Gate donation $1.00. Camping $3.00 per night; $5.00/2 nights. Talk in on 146.22/82 repeater. For information: Ron Dishman, N1CMZ, 37 Marlboro Avenue, Augusta, ME 04330. (207) 623-8351.

NEW JERSEY: The Ramapo Mountain ARC, WA2SNA, presents its 9th annual flea market, August 17, Oakland American Legion Hall, 65 Oak Street, Oakland, 20 miles from GW bridge. Admission $1.00. Non-ham family members free. Indoor tables $6.50. Tailgating $3.00. Talk in on 147.49/146.49 and 52. For information: Tom Risseeuw, N2AAZ, 63 Page Drive, Oakland, NJ 07436. Tel. 337-8389 after 6 PM.

WISCONSIN: Green Bay Mike & Key Club's Summer Swapfest, Saturday, August 17, Ashwaubenon Community Center, Anderson Drive across from Baypark Square Mall. Free admission and parking. Doors open 8 AM. Sellers 7 AM. Buy, sell, trade. Reserved 8' tables $5.00. Limit 4. SASE with check to Green Bay Mike & Key Club, Bill Johnson, N9CNO, 2177 Orrie Lane, Green Bay, WI 54304. (414) 494-8948.

MISSOURI: The Ozarks Amateur Radio Society's 4th annual Congress & Swapfest, Sunday, September 8, City Park, Jct. of US 60 and Highway 37, Monett. Swapfest 11:00 AM. Buffet dinner 1:00 PM. No tickets necessary. All Amateurs and families welcome. Talk in on 146.37/.97, 146.52 and 7.250 MHz. For information: Ozarks Amateur Radio Society, Box 327, Aurora, MO 65605. (417) 678-5330.

GEORGIA: Augusta Hamfest, September 15. Dealers and tailgaters welcome. Food and drinks available. ARRL/VEC exams 8 AM. Tickets $1.00. 6/$5, 13/$10. Talk in on 34/94. SASE to Bill Hardin, 4430 Forrest Drive, Martinez, GA 30907. (404) 863-4360.

INDIANA: The Tippecanoe Amateur Radio Association's 14th annual Hamfest, Sunday, August 18, Tippecanoe County Fairgrounds, Teal Road and 18th Street, Lafayette. Grounds open 7 AM. Tickets $3.00. Large flea market, dealers, refreshments and fun. Talk in on 13/73 or 52. For tickets or information: Lafayette Hamfest, Route 1, Box 63, West Point, IN 47992.

ILLINOIS: Bolingbrook Amateur Radio Society's B.A.R.S. Hamfest '85, Sunday, September 8, Santa Fe Park, 91st and Wolf Road, Willow Springs. Advanced registration $2.00. $3.00 at the gate. Talk in on 147.33/93 and 146.52. For information: Ed Weinstein, WD9AYR, 7511 Walnut Ave., Woodridge, IL 60517. (312) 985-0527.

MICHIGAN: The Grand Rapids Amateur Radio Association's annual Swap and Shop. Saturday, September 21, Hudsonville Fairgrounds. Dealers, indoor sales area, outdoor trunk swap, concession. Gates open 8 AM. Talk in on 146.16/76. For information: Grand Rapids ARA, PO Box 1248, Grand Rapids, MI 49501.

PENNSYLVANIA: The Skyview Radio Society's annual Hamfest, Sunday, September 16, Clu Grounds, Turkey Ridge Rd., New Kensington. Noon to 4 PM. Registration $2.00. Vendors $4.00. Talk in on 146.04-.64 and 52.

PENNSYLVANIA: The Central Pennsylvania Repeater Association's 12th annual Hamfest/Computerfest, August 25, adjacent to Hersheypark, Chocolate Town, USA. Registration $3.00. Children 12 and under free. Special reduced admission to Hersheypark for registrants and families. Large indoor dealer and flea market. Large outdoor tailgate area. Food and refreshments. Talk in on 145.47 repeater or 146.52 simplex WA3KXG. For information: Paul W. McDonnell, N3BKI, (717) 697-1880, noon to 8 PM.

PENNSYLVANIA: The Uniontown Amateur Radio Club (W3PIE) will hold its 36th annual Gabfest, Saturday, September 7, Club grounds, Old Pittsburgh Road, Uniontown. Registration $3.00 or 2/$5.00. Free Parking — Free Coffee — Free Swap & Shop with registration. Talk in on 147.645-.045 & 144.57-.17. For information: John Cermak, WB3DOD, U.A.R.C. Gabfest Committee, PO Box 433, Republic, PA 15475. (412) 246-2870.

VIRGINIA/WEST VIRGINIA: The Bluefield Hamfest, sponsored by the East River Amateur Radio Club, will be held Sunday, August 25, Brushfork Armory Civic Center, 1 mile north of Bluefield, WV. 9 AM to 3 PM. Admission $4.00. Children under 12 free. Large indoor flea market, satellite TV and various specialty dealers. Paved parking, food on site, other activities. Walk in license exams 9 AM. Bring copy of license and completed 610 Form. $4.00 fee. Talk in on 144.89/145.49 and 146.52. For information: Jim Perdue, KC8NG, Rt. 5, Box 457, Bluefield, WV 24701.

NEW YORK: The Putnam Emergency Amateur League (PEARL) will have its annual Electronics Extravaganza, August 17, 9 AM to 4 PM, J.F. Kennedy Elementary School, Brewster. Admission $2.00. Tables $5.00. Walk in VEC exams. For table reservations and information: R. Dillion, N2EFA, RFD 7, Noel Court, Brewster, NY 10509. Talk in on 144.535/145.135.

ALABAMA: The Huntsville Hamfest, Saturday and Sunday, August 17 and 18, Von Braun Civic Center in Huntsville. Free admission. Exhibits, forums, air-conditioned indoor flea market and non ham activities. Walk in FCC exams 9 AM Saturday, August 17. Family tours of the Alabama Space & Rocket Center available. Some camp sites with hookups available, first come, first served. Reserved flea market tables $5/day. For more information: Huntsville Hamfest, 2804 S. Memorial Parkway, Huntsville, AL 35801.

PENNSYLVANIA: The Mid Atantic Amateur Radio Club's annual Hamfest, Sunday, August 11, 9 AM to 4 PM, rain or shine, Bucks County Drive-In Theater, Rt. 611, Warrington. Admission $3.00 + $2.00 for tailgating. Setup starts 8 AM. Bring your own table. Plenty of parking, refreshments. Talk in on WB3JOE/R, 147.66/.06 or 146.52. For information: MARC, PO Box 352, Villanova, PA 19085 or call Bob, WA3PZO (215) 449-9727.

ARKANSAS: The 16th annual Queen Wilhelmina Hamfest, Queen Wilhelmina State Park, September 7 and 8. This beautiful state park facility on top of Rich Mountain near Mena, offers family fun and relaxation. Free admission, dealer display, Saturday night banquet, $7.00, camping, tailgating, flea market, miniature golf, wild life zoo, new playground, miniature train ride, ladies' tour. Talk in on 146.19/.79. For information: John Harris, KC5XK, 5018 S. 9th, Ft. Smith AR 72903.

ILLINOIS: Vermilion County Hamfest, August 25, W9MJL Clubhouse, Harrison Park West, Danville. Donation $1.50 at gate; $1.00 advance. Saturday evening steak cookout, $5.00/reservations. Talk in on 146.22-82. For information and reservations: Joe Mayer, KB9GS, 613 E. Kelly, Box 356, Westville, IL (217) 267-2946.

OPERATING EVENTS

"Things to do..."

Riding Radio Operators — Amateur Radio Motorcycle Club Net meets every Thursday night at 0300 UTC at 3888 kHz standard time and 7237.5 kHz daylight saving time. An eastern USA group meets one hour earlier at 3888 kHz year-round. Send business SASE to AG0N, Gary McDuffie, Rt. 1, Box 464, Bayard, NE 69334 and ask for net information.

September 4: Howdy Days. Eligibility — all licensed women operators throughout the world. Operations — all bands and modes. No cross band. Station counted only once. Exchange — YLRL member or non YLRL member. Score 2 points for each YLRL member worked and 1 for non YLRL member. All logs must show YLRL membership or not, score and must be received by October 4, 1985. Send to Marty Silver, NY4H, 3118 Eton Road, Raleigh, NC 27608.

August 18: The DuPage Amateur Radio Club will operate special event station W9DUP from the War Museum submarine, U.S.S. Silversides, Navy Pier in Chicago. 1300Z August 18 to 0200Z August 19. For a special submarine QSL card, SASE to W9DUP, PO Box 71, Clarendon Hills, IL 60514.

August 17: 26th annual New Jersey QSO Party: 2000 UTC Saturday, August 17 to 0700 UTC Sunday, August 18 and from 1300 UTC Sunday, August 18 to 0200 UTC Monday, August 19. Phone and CW are considered same contest but separate bands. Station may be contacted once on each band. Suggest phone activity on even hours; 15 meters on odd hours. Exchange QSO number, RST, and QTH. NJ stations send county for QTH. Send logs and comments to: Englewood Amateur Radio Association, PO Box 528, Englewood, NJ 07631-0528. Include #10 SASE for results.

THE GUERRI REPORT

Ernie Guerri W6MGI

surface-mounted components improve circuit designs

The traditional approach to printed circuit board design has been to lay out the board so that component leads go through holes in the board and are soldered to pads on one or both sides of the board. A technique first developed in the mid-1970's, using Surface Mounted Components (SMCs), has finally caught on, and achieves dramatic improvement over standard PCB designs. In the new approach, the PCB is designed for the same functional application, but the components are all mounted directly on the surface of the circuit paths themselves — *no holes*! All of the components are designed with short, flat leads that are flush with the PC path. This technique is particularly useful for RF circuits because lead lengths are dramatically shortened, thereby reducing parasitic inductance and capacitance, and improving EMI/RFI problems. Proper design of the components can maintain the impedance integrity of stripline designs right into the active region of the component. Circuits using this approach at 1200 to 1500 MHz give performance nearly as good as their low frequency counterparts. Let's hope that this attracts some real attention to the development of more equipment for the 1200 MHz Amateur band! Additionally, manufacturing costs are significantly reduced and reliability is improved. Circuit densities can be increased by 30 to 50 percent, resulting in considerable reduction in size, for use in complex equipment. The technique is particularly well suited to full automated production, and high volume producers such as the audio and TV industry are now regularly delivering products incorporating this approach. We should see the first uses of this improved technique in Amateur equipment in the immediate future.

computer-aided everything

The flood of information that seems to fill everyone's mailbox these days includes an increasing amount of data promoting the ways in which a computer can make each of our personal endeavors a snap. Productivity is the magic word, but much of what's offered seems to actually add complexity to such nominally simple tasks as home budget management.

This is not the case with computer-aided engineering, design, and manufacturing — CAE, CAD, CAM as they're called. Each of these tasks normally involves thousands of steps, each of which must be executed exactly, in complete compliance with design rules, and in concert with other phases of the process. All must be organized so that the final product is technically and physically correct, and on time. Modern product development cycles are frequently so short that there would be no way to do all the necessary steps by hand.

Take the example of designing a new computer chip. Such a chip may have a complex architecture, and 30 to 40 thousand active elements. By using computer-aided techniques, the design rules for circuit interconnects, layout, propagation delays, etc., can all be simultaneously considered every time a single change is made. A CAE tool called a "silicon compiler" actually contains the design rules for making the IC masks stored in computer memories. As the system architect and design engineer work on the chip's functional characteristics, the silicon compiler automatically includes the necessary semiconductors, routings, chip real-estate and thermal characteristics in the final design. Using these techniques, three employees of a major computer manufacturer were able to design a 32-bit minicomputer chip set (including nearly 40,000 transistors) in less than a year.

It's been estimated that there are fewer than 5000 integrated circuit designers in the entire world. With the design tools offered by CAE and CAD systems, at least half of all electronic engineers could participate directly in the design of semi-custom ICs.

The prospects for even more dramatic advances in electronics are thus enhanced by the broad participation of another half-million or so engineers whose creativity is now frustrated.

microwaves cook rocks

The success of commercial mining operations depends on assessing the extent to which rocks bearing sufficient quantities of desired material can be processed. Therefore, the quantity of desired ore per ton of rock is an important measure of the ultimate financial value of the process. Because one of the major problems in ore processing is keeping impurities out of the desired product, complex — and therefore expensive — steps must be taken to resolve these problems.

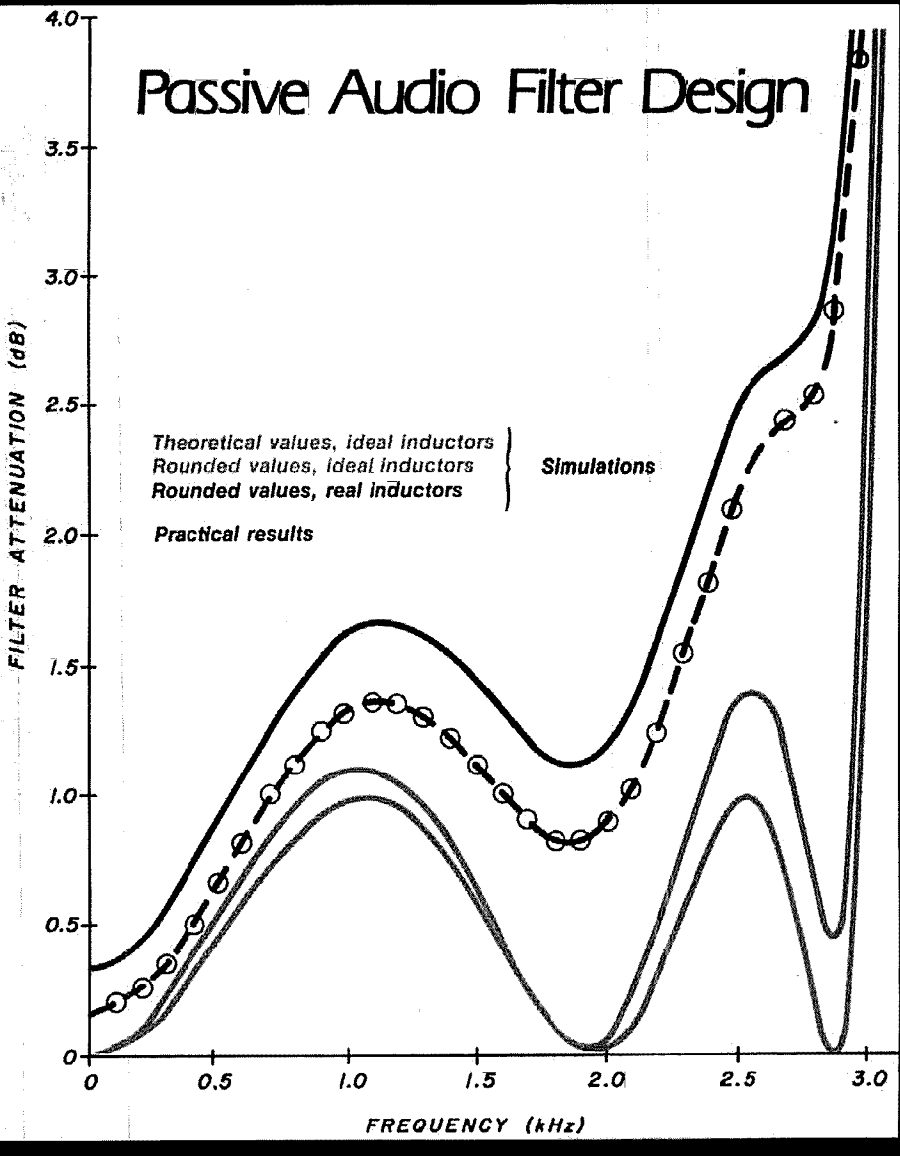
A new technique developed by a Colorado company may represent a major breakthrough in this field. Recognizing that each element has an atomic structure that can be excited by external energy sources, the company has devised a technique for illuminating rocks containing various elements with microwave energy matched to the resonant frequency of the element. The result is an ability to selectively melt desired material and leave surrounding material and impurities "cool" and undisturbed. Although the energy expenditure/recovery ratio is not obvious, the technique merits watching as yet one more example of RF in the "workhorse" environment.

ham radio

SEPTEMBER 1985 / $2.50

new column: PRACTICALLY SPEAKING by Joe Carr, K4IPV

ham radio magazine



hr focus on communications technology

- adjusting SSB amplifiers
- understanding telephones
- a DTMF controller for repeaters
- a look at probes
- digital satellite tracker
- carrier-activated CW reception limiter
- simple continuity tester
- plus W9JUV, W6SAI, W1JR, W6MGI, K0RYW

ham radio magazine

SEPTEMBER 1985

volume 18, number 9

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contents

REFLECTIONS

it can happen to anyone

One of the more pleasurable obligations of working in the Amateur Radio industry is keeping up-to-date on all the latest equipment, accessories, and doo-dads that come along.

One night not too long ago I took one of the newest radios home to do a little informal testing. Like most radios on the market today, this one is loaded with just about every feature you'd want to have. In fact, if it had legs, I guess it would walk the dog in the morning — if I remembered to set the timer the night before. After making up all the cables and installing the radio, I sit down to make its acquaintance. All goes well for the first few minutes . . . signals are strong and the radio operates flawlessly. Until, that is, I decide to QSY to the broadcast band to listen to the local news.

The radio goes dead.

My first reaction is to assume I'd either turned the wrong knob or pushed the wrong button, so I run through the standard drill of trouble-shooting: punch every knob and turn every button to see if *that* solves the problem. Five minutes of this pass, and the problem is still there . . . the radio is *dead*.

Now what? "Back to the manufacturer," I sigh, and off I go to bed. It's late, and I'm completely discouraged. "Why can't they build something that *works*?" I complain to my sleeping XYL. I am utterly discouraged.

After I turn out the light, and I'm lying there waiting to fall asleep, I ask myself, "Did I try the squelch?" I'm not sure, so I throw on a robe and head for the door. My wife stirs in her sleep, pulls the covers to *her* side, and murmurs something good-naturedly unintelligible. I run downstairs, turn on the radio, set it to the broadcast band, and wait for the problem to reappear. It does. I turn off the squelch.

BINGO.

The bottom line, I guess, is that problems like this can strike us all. From Novice to Extra, no one is immune. The increasing complexity of our world — and of Amateur Radio — makes it hard to be fully expert on the many different pieces of equipment and in the varied technologies we use. So when a problem occurs, all we can do is walk through a step-by-step check of each and every control to determine whether it's really an equipment problem or just another embarrassing incidence of "cockpit error." If the malfunction escapes this scrutiny, then we can consult the owner's manual to see if the problem is one that the manufacturer has already anticipated. If this step fails, we can check with the dealer to see what he has to say. (Most likely, he'll be thoroughly equipped to handle the problem.)

My point is that sending equipment back to the manufacturer is always the action of last resort. Yes, the manufacturer has a skilled service staff to troubleshoot and repair equipment. But as Joe Carr, K4IPV, points out in his new *ham radio* column (see page 67), as much as 40 percent of the equipment that lands on the benches of professional service technicians has absolutely nothing wrong with it! By not taking time to really check things out first, its owners are inconvenienced, off the air, and out the cost of packing and postage.

These newfangled radios sure are nice. But they're a whole different world from the Hammarlund HQ-129 and Johnson Ranger I started out with.

Craig Clark, N1ACH
Assistant Publisher

presstop de W9JUV

AMATEUR USE OF A GEOSTATIONARY SATELLITE has been proposed by NASA. One of NASA's Advanced Communications Technology Satellites (ACTS) is to be made available for use by experimenters for a two-year period starting in 1989; Amateur Radio is one of the groups that NASA expects to be interested in using it. Uplink band will be 27.5-30 GHz; downlink, 17.7-20.2 GHz. As these are outside the Amateur bands, it's likely either an STA or experimental license will be required for them. The satellite will use highly directive antennas aimed at some major population centers, but NASA even suggests Amateurs might use repeater techniques to permit others outside the bird's antenna "footprint" access to it. Interested Amateurs should request NASA's November 1984 "Notice of Intent" from Ron Schertler, ACTS Projects Experiment Manager, MS-54-6, NASA, 21000 Brookpark Road, Cleveland, Ohio 44135.

BROADCASTERS MAY DO ANYTHING THEY WISH WITH AMATEUR TRANSMISSIONS, but Amateurs are strictly forbidden to use Amateur Radio in any way that might be construed to benefit any "broadcaster." The FCC's Report and Order on BC Docket 79-47 now appears to almost forbid such well established Amateur Radio "public service" activities as, for example, any direct tie between an Amateur Radio weather net and the U.S. Weather Bureau, since Weather Bureau information is "broadcast" on NOAA's 162-MHz weather stations.

The Redefinition Of "Emergency" in Part 97 Is At The Heart of the problem. Verified emergency communications such as "The dam has broken" or "A tornado just crossed the river" would still be permitted, of course. What appears to be against the new rules is letting an Amateur station in a weather net tell the Weather Bureau "The dam looks shaky" or "The sky's getting very dark." Under the new rules, however, nothing would prevent the Weather Bureau, broadcasters, or anyone else from monitoring an Amateur Radio weather net and using the reports heard in any way they wished.

A Petition For Reconsideration Has Been Filed By ARRL, asking the Commission to go back to the former less stringent definition of "Emergency," or to temper the new one so it won't preclude public service ties between weather nets and the Weather Bureau.

AUXILIARY OPERATION ON ALL AMATEUR FREQUENCIES except 431-433 MHz and 435-438 MHz has been proposed by the Commission in PR Docket 85-215. The present rules restrict auxiliary operation (principally control links) to above 220.5 MHz. However, the FCC feels that there is no reason, given the present state of the art, that such auxiliary operation shouldn't be permitted on the other Amateur bands, as well.

Comments On PR Docket 85-215 Are Due At The Commission September 24, and Reply Comments by October 25.

NEVADA HAS JOINED THE SHIFT TO 20 KHZ 2-METER SPACING, except for the Las Vegas area, which has traditionally gone along with Southern California. Eastern Nevada will be coordinated by the Utah coordinator, and the western part by NARC's northern California coordinator. Iowa's council, however, has voted to remain at 15 kHz and the ARRL bandplan though with a voluntary state-wide tone access system. They also intend shifting the 145-MHz repeater subband to 15 kHz spacing in order to pick up additional channels.

800-MHZ SCANNERS WOULD BE OUTLAWED IN CALIFORNIA under a proposal, SB 1431, presently being considered in the state legislature. The proposal, apparently the result of phone company concern over the security of cellular communications, has California Amateurs worried over possible negative impact on Amateurs. Los Angeles attorney Joe Merdler, N6AHU, has been instrumental in delaying the bill's approval, though scanner manufacturers haven't seemed very concerned about it. If it becomes law a court challenge seems likely.

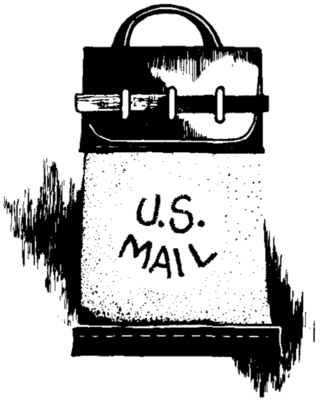
AN 800 NUMBER FOR QUESTIONS ABOUT ITS VEC PROGRAM, including its own or any other VEC's exam schedule, has been installed by DeVry. The number, which is presently accessible only in the greater Midwest area, is 1-800-327-2444; A DeVry WATS line covering the entire Continental U.S. will be announced in the near future.

"The Volunteer Examiner," A VE Newsletter created and edited by DeVry VE KI9R, is now being distributed by DeVry. An SASE to Jim Georgius, W9JUG, DeVry ARS, 3300 N. Campbell, Chicago, Illinois 60618 will bring a sample copy.

W9JUG's Sunday VE Net Now Meets On 7173 kHz, 1400Z, with a second session at 1800Z.

"12-12 WORLDWIDE" IS A NEWLY FORMED ORGANIZATION to promote activities on the new 24 MHz (12 meter) WARC band. For details send an SASE to Steve Walz, WA5UTO, Box 222, 318 S. Massachusetts Avenue, Cherokee, Oklahoma 73728, or call him at (405) 596-3487.

AN FCC CITATION FOR "BROADCASTING" HAS COST K6KPS $2000. K6KPS, James Brantley, has been a problem to 20-meter phone band users for decades, calling CQ but ignoring responses and carrying on imaginary QSOs on top of various nets, DX pileups, phone patches, and ordinary rag chews with a big signal from the Los Angeles area. Though he's done this for years (K6KPS often broke up a 14295 "commuter's net"—in which Presstop editor W9JUV participated in the mid-60s) and has been the subject of previous FCC investigations, the fine was levied for his operating tactics observed during last December.



comments

new band privileges for Novice operators

Dear HR:

I was amazed and very pleased to see your July *Presstop* item about the proposed upgrading of Novice privileges. I have been a Novice ham operator since 1979 and am pleased to see the prospect of a reward in the form of more privileges being considered for the Novice class license.

I would like to encourage all Novice operators who read *ham radio* to speak out and let their feelings on this subject be known to the ARRL and the FCC. This may be our only chance to assist in changing the long-time rules and regulations regarding our hobby. It may very well be a turning point that will create the interest to save the hobby as we know it.

Allowing the Novice use of phone on 10 meters, 220 MHz, and 23 cm can only help to encourage us to strive for the privileges of the higher grade licenses. CW is an entertaining and rewarding mode of communication, but I believe the Novice license should give more to encourage us to seek out our full potential as operators. A simple pat-on-the-back in the form of a reward would be all it would take for many of us.

Technicians would, of course, benefit from these new privileges also by allowing phone on 10 meters. I might go even a step further and suggest not only the use of SSB, but even FM for both license grades.

I do not believe for a minute that this rule change would degrade the bands with "glorified CB operators" because the license elements are still sufficiently difficult to discourage such abuse. Besides, that era was just a passing fancy for many.

With today's increasing use of digital communications, the new rules would also allow Novices to utilize the state-of-the-art computer systems to further their knowledge of present-day communications.

All in all, I think it would be foolish to not allow us the opportunity to take part in this ever growing advancement in technology.

T. Dillingham, KAØDOE
Aspen, Colorado

used equipment for new hams

Dear HR:

Much has been said recently about bringing new blood into the Amateur Radio hobby, and several articles have focused on the wonderful services being provided in the public interest, convenience, or necessity. All of these remarks are laudable, and the two situations can easily go hand-in-hand. Additionally, a few other problems we must each face someday can easily be solved now, while helping with the first two I mentioned: what will become of our equipment when we tire of it, or leave for the land of the silent key? And how can we find replacements for the tax loopholes about the run out of town on the Amtrak rails? (These remarks are addressed to dealers as well as the average ham.)

I suppose we have all seen on the "used" shelf at the few remaining Amateur Radio stores some piece of equipment that could be put to better use in a shack somewhere. Perhaps we have a piece or two in our attic or garage that we haven't really been using in the recent past, but hesitate to just throw away. Maybe one of our senior citizens has some neat stuff that his or her surviving spouse will have some problem with disposing of in some coming period of distress. One cure would be to put that gear to good use now, getting the tax advantage now, providing a service now, and perhaps helping out some youngster getting started down that primrose path we have each delighted in following.

There are literally thousands of applications for used equipment in each of our hometown areas, and many of them are equally deserving of attention; there is simply no room here to list them all, but a few will get your minds moving in the right direction. Rigs and associated equipment of appurtenances are sorely needed by schools, camps, Courage Handi-hams, Civil Defense, FEMA, DSA, National Weather Service, Fire and Police Departments, and our many clubs with active Novice classes, among countless others. Perhaps your donation of usable and serviceable equipment now will be just what is needed to get new hams on line or to provide that equipment needed by present-day hams to provide that vital service to the public that will keep this hobby active and thriving.

Think about it now, while you can still do something about it, and before the spouse cleans house for you, or a good piece of equipment self-destructs on your shelf!

Jerry Murphy, K8YUW
Lakewood, Ohio

off and on

I recently saw a 12 VDC to 110 VAC power inverter that was switched on and off on demand by the load, without extra switches or wires.

Do any of your readers have information on how this is done?

Sam Popkin, K2DNR/7
Mesa, Arizona

passive audio filter design
part 1: development and analysis

Circuits using miniature preferred-value components are adaptable to printed-circuit boards

A recent article in *ham radio*[1] was the latest of a long series by W3NQN on the subject of elliptic lowpass audio filters suitable for Amateur constructors. It is mainly through the efforts of W3NQN that these filters have remained in the Amateur literature despite the proliferation of designs for active audio filters using operational amplifiers.[2-5]

It is surprising that active filters have become so popular considering the complexity required to equal the performance of passive designs and the disadvantages associated with active elements. They require power, produce noise, have a limited ability to handle large signals and have a limited upper operating frequency unless expensive, high-frequency devices are used. These disadvantages usually don't apply when using a passive circuit containing inductors.

Probably the main reasons for the decline of passive filters are the misunderstandings surrounding the selection of components for them. Some of these misunderstandings can be explained by:

- The lack of published designs using preferred-value inductors and capacitors.
- The belief that highly accurate inductance and capacitance values must be used to obtain acceptable results.
- The belief that only high-Q inductors are suitable for use in filters.
- Ignorance of the fact that ready-wound, standard-value inductors are available easily and inexpensively.
- A general phobia of winding inductors, especially ones with several hundred turns of wire.

W3NQN has avoided these problems by using telephone-line loading coils in his designs. These loading coils are relatively high-Q components. Specialized designs have been developed which make use of the inductance values included in these coils, or by slight modifications to the inductors by removing turns.[6]

In general, the capacitance values required in a passive filter are non-preferred and are obtained either by connecting several preferred value capacitors in parallel or by measuring capacitors and selecting the required value. Designs have been published in which some[7,8] or all[9,10] of the capacitors are preferred value, but these are intended for RF applications in which the non-preferred value inductors are not too difficult to wind and will have a relatively high Q value.

The question of what effect the inductor Q has on a filter response does not seem to have been addressed in the United States Amateur literature. Only two references that consider this question could be found,[11,11A] but non-preferred values of capacitance and inductance were used. The results from this investigation were encouraging, with only a slight degradation in the flatness of the passband and rounding in the cut off region of the Butterworth, Chebyshev, and Bessel filters considered.

objectives

One aim of this series of articles is to describe my investigations into the use of miniature preferred value inductors and capacitors in several types of audio filter. My goal was to use a single, unselected, inexpensive component for each inductance and capacitance value so that passive filter construction (and design, if the reader so wishes) can be simple for even the most inexperienced Amateur. Filters using these components are extremely compact and suitable for construction on printed circuit boards in modern equipment — which, unfortunately, designs using the telephone line loading coils are not.

The investigations were carried out using computer

By Stefan Niewiadomski, 29 MacKinley Avenue, Stapleford, Nottingham, NG9 8HU, England

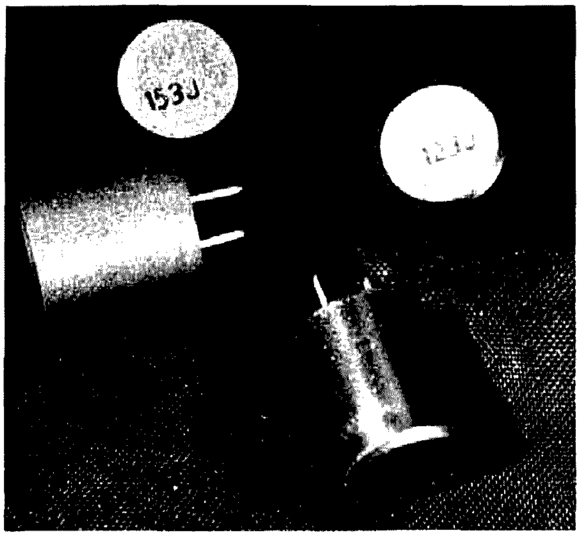


fig. 1. The Toko inductors used in the author's filter studies.

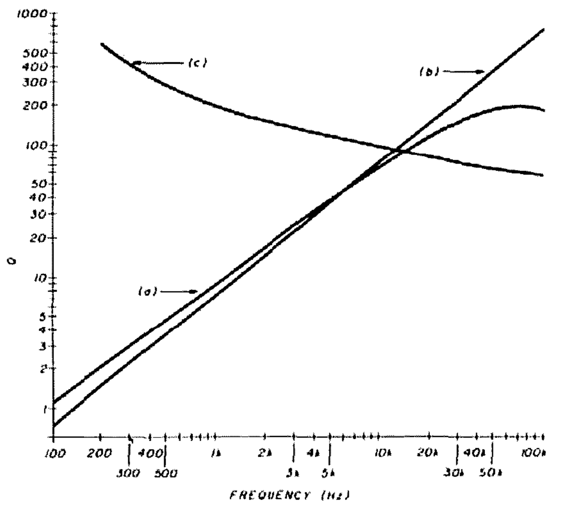


fig. 2. Plot of Q versus frequency for (*a*) 22-mH Toko 10RB inductor; (*b*) ideal 27-mH inductor and 22 ohm resistor; and (*c*) Siemens 0.22-μF capacitor.

simulations and practical tests. The simulations were performed on an Apple II microcomputer, using a BASIC circuit analysis program.[12] Circuit simulation programs that run on several home computers are now available.*

The major advantage of using circuit simulation in this field is that the effects of altering component values and using low-Q inductors can be isolated from each other and investigated separately. This contrasts with an experimental approach, in which the combined effects of non-exact component values and low-Q inductors would have to be accepted unless a large number of inductors of various values and Qs could be wound — which is obviously impractical for the Amateur.

I will also discuss attenuation equalizers and show how these comparatively simple circuits can improve the responses of filters made from low-Q inductors. I believe this is the first time these networks have been described in a practical way in the Amateur literature.

Also discussed in this series of articles are some methods of incorporating an audio filter into an audio path to ensure correct matching of the filter. Other topics tackled are the question of how much passband ripple is acceptable in an audio filter used in an Amateur receiver, and how passband ripple and stop-band attenuation can be traded off to produce improved overall performance.

components

The inductors used in this study are the Toko 10RB range, manufactured from 1 mH to 120 mH with 5 percent tolerance. Another range, the 10RBH, is available from 150 mH to 1500 mH with 10 percent tolerance.† These two inductor types have identical case styles. They are cylindrical with a diameter of 10.5 mm and a height of 14 mm (see **fig. 1**). Leads are radial with a 5-mm pitch, enabling an extremely compact PCB layout.

Examination of a 47-mH inductor after removing its outer ferrite casing revealed a wire size of approximately No. 40 AWG. For most of the 10RB range, the Q quoted[13] is a minimum of 100 at a test frequency of 50 kHz. The recommended operating frequency range for these coils is from 100 kHz to approximately 170 kHz[14] over which they have a sufficiently high Q to be considered as almost ideal inductances.

The Q at audio frequencies is much lower and was measured for a 27-mH coil from 100 Hz to over 100 kHz using a Hewlett Packard 4192A LF Impedance Analyzer. **Figure 2** shows the results obtained. At 100 Hz, the Q is 1.1, rising to about 9 at 1 kHz, 24 at 3 kHz, and more than 50 at 7 kHz. At frequencies above 80 kHz, the Q begins to tail off, as the maximum operating frequency of the ferrite is approached and skin effects in the wire become appreciable.

Also shown in **fig. 2** is a plot of the Q of an ideal inductor of 27 mH with a 22-ohm resistor in series, which is the quoted DC resistance of the 27-mH Toko inductor. This model seems to be accurate at about

*One available in the UK is sold by Number One Systems, 9A Crown Street, St. Ives, Huntingdon, Cambs, England. Versions are available for the BBC model B and Sinclair Spectrum 48K computers.

†To order, contact Toko America, Inc., 1250 Feehanville Drive, Mt. Prospect, Illinois 60056 (312 297-0070).

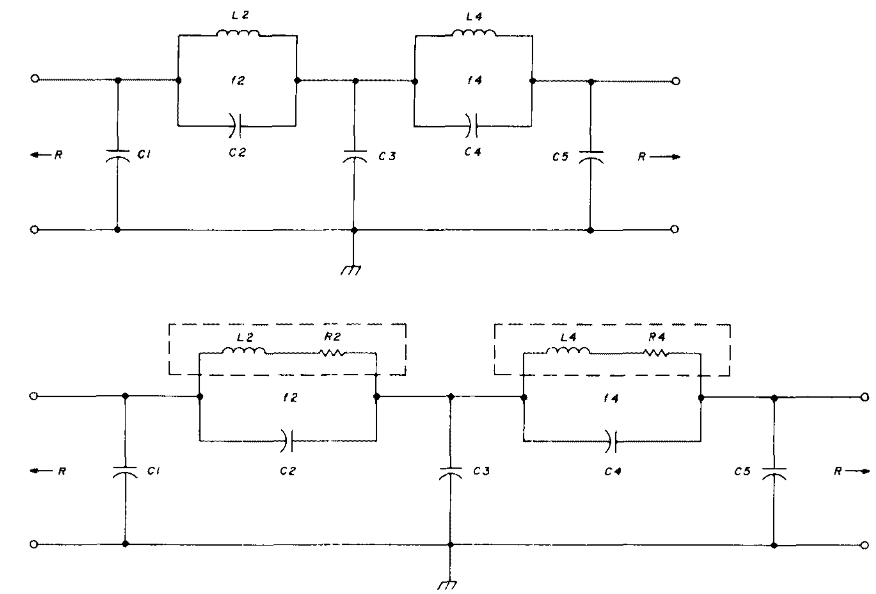


fig. 3. Elliptic 5-branch low-pass filter. (*A*) Schematic of filter with ideal inductors, and (*B*) schematic of filter with real inductors.

6 kHz, giving worse results below 6 kHz and better results above 6 kHz.

Ideally, to accurately model these inductors, a plot of Q versus frequency should be obtained for each value of inductor and the series resistance determined which most closely follows this plot. For the simulations presented here, the inductors were modeled with a series resistor equal to the manufacturer's quoted DC resistance, considering that this will give slightly worse results than in practice.

The DC current flowing through these coils, particularly in the 10RBH range, should be kept to a minimum to avoid heating and core saturation. In fact, to be completely safe, it is better to keep the DC current at zero. In most applications, this is achieved by inserting a DC blocking capacitor in series with the input of the filter in which the inductors are used when there is a direct DC path to a potential different from the input source. This also has the advantage of obviating the danger of interaction between the DC conditions of the driving and terminating circuits. In most applications a capacitor of 100 μF will avoid modifying the response of a filter, but the value can be chosen to have beneficial effects, which will be discussed later.

There are many types of capacitors suitable for use in audio filters. The ones I prefer are the Siemens B32560/1/2 metalized polyester layer type, which are available in a wide range of values and are very compact for their capacitance*. The variation of Q with

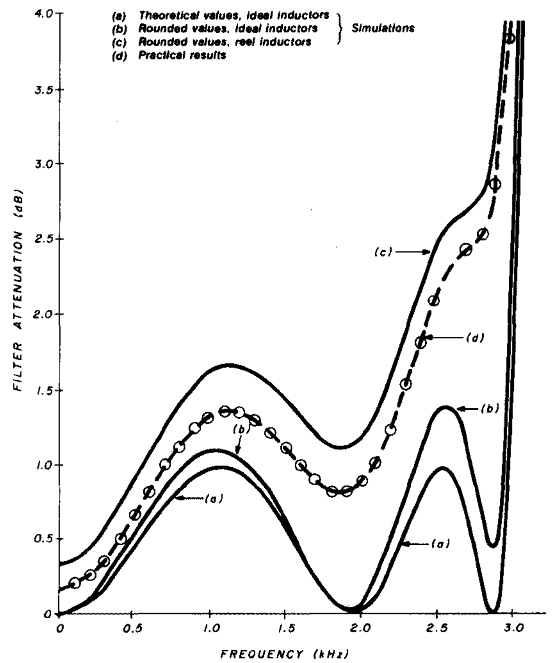


fig. 4. Passband response of 1-dB/50-dB 5-branch 3-kHz elliptic low-pass filter, showing simulation and practical results.

*Panasonic capacitors, available from Digi-Key Corp., P.O. Box 677, Thief River Falls, Minnesota 56701, (1-800-344-4539) may be substituted. —Ed.

frequency of a 0.22μF B32560 capacitor is also shown in **fig. 2**. This capacitor maintains a Q of more than 100 at frequencies up to 10 kHz and can be accurately modeled as a pure capacitance at audio frequencies. These capacitors can be obtained with a 5-percent tolerance and 7.5-mm lead pitch. These are used here.

low-pass filters

The most popular filter for audio low-pass applications is the elliptic or Cauer filter, named for the German network theorist who developed the mathematics associated with its characteristics. An excellent introduction to elliptic filters is given in reference 15, but many of W3NQN's articles include descriptions of the terminology of these filters and their fundamental characteristics. A brief resume of these points is given in the **Appendix**.

The first design considered here is the five-branch low-pass whose circuit is shown in **fig. 3**. **Figure 3A** shows ideal inductors and **fig. 3B** shows a resistor in series with each inductor to simulate their losses. The specification for this filter is:

passband ripple (A_p)	1dB
stopband minimum attenuation (A_s)	50 dB
ripple cutoff frequency (f_{A_p})	3000 Hz
start of stopband frequency (f_{A_s})	4221 Hz
source impedance	500 ohms
load impedance	500 ohms

From now on, I will refer to this filter as the 1-dB/50-dB filter, indicating its passband ripple and minimum stopband attenuation; this terminology will also be used for the other low-pass filters described later.

Table 1 shows the component values for the 1-ohm, 1-rad/sec prototype,[16] the exact values for the filter scaled to 500 ohms and 3 kHz, and the rounded values both with and without resistors to simulate the lossy real inductors. Each capacitor and inductor has simply been rounded to its nearest E12 series preferred value.*

The simulation and practical results for this filter are shown in **figs. 4, 5,** and **6**. Three sets of graphs have been plotted so that the axes for each set can be chosen to show the maximum detail in the passband,

*Capacitors are commercially available in several series of preferred values, two of which have designations of E12 (with a 10 percent tolerance) and E24 (with a 5 percent tolerance), (16A). The reciprocal of the E-number is the power to which 10 is raised to give the step multiplier for that particular series. For example, for the E12 series, the step multiplier is 10^(1/12) or 10^0.083333 = 1.2115. The sequence of the series is therefore 10, 10 • 1.2115 = 12, 12 • 1.2115 = 15, 15 • 1.2115 = 18, 18 • 1.2115 = 22, 22 + 1.2115 = 27, etc. — **Ed.**

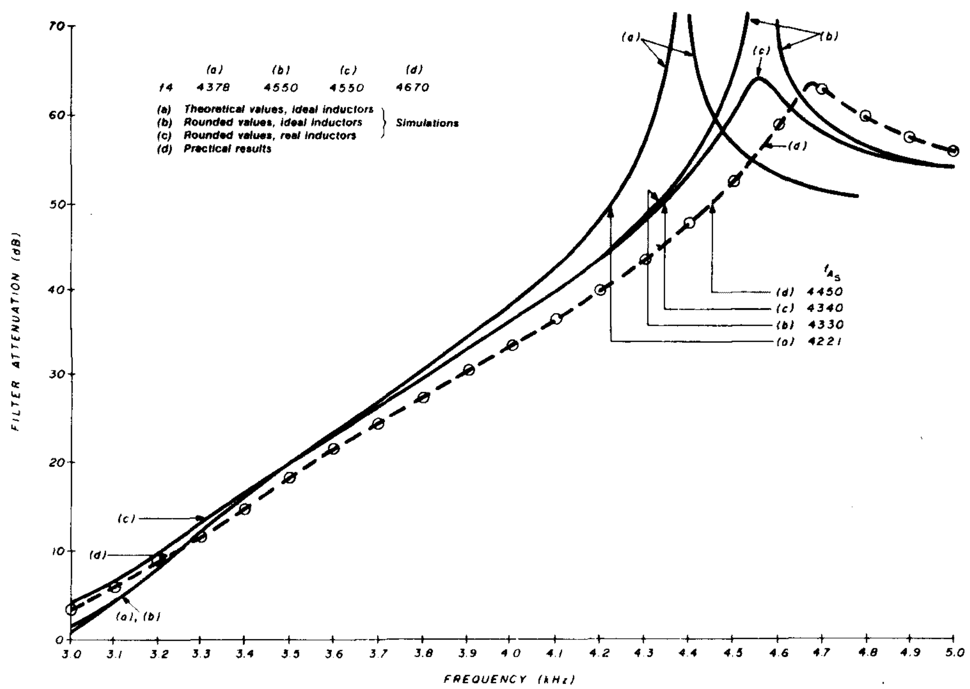


fig. 5. Transition band (and part of stopband) response of 1-dB/50-dB 5-branch 3-kHz elliptic low-pass filter, showing simulation and practical results.

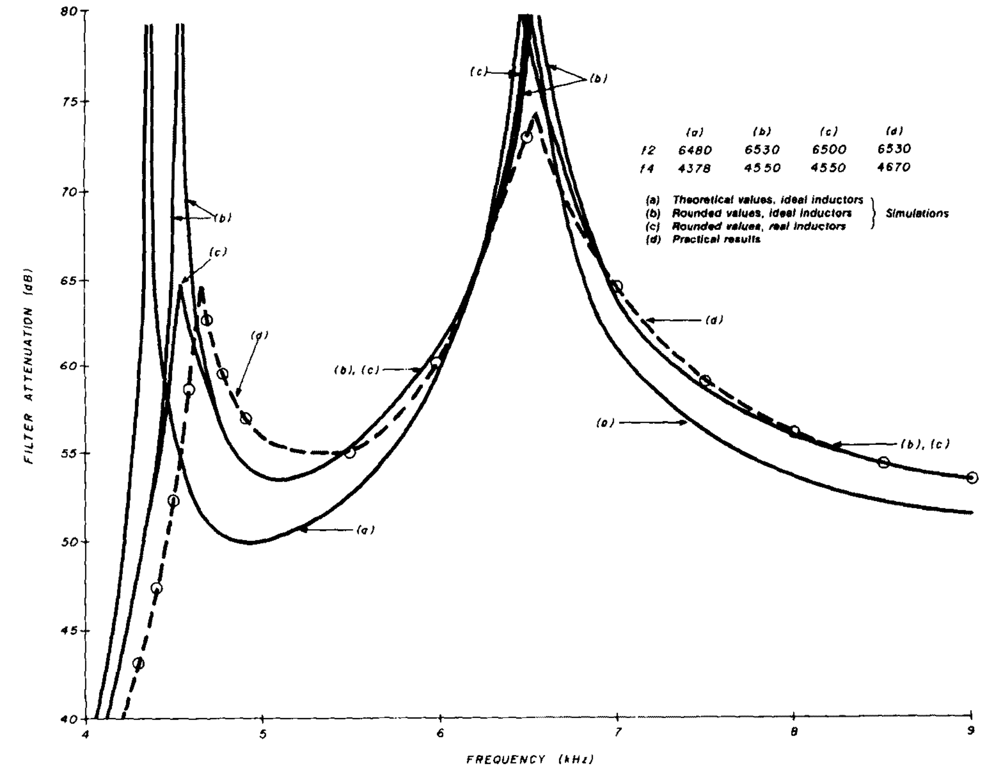
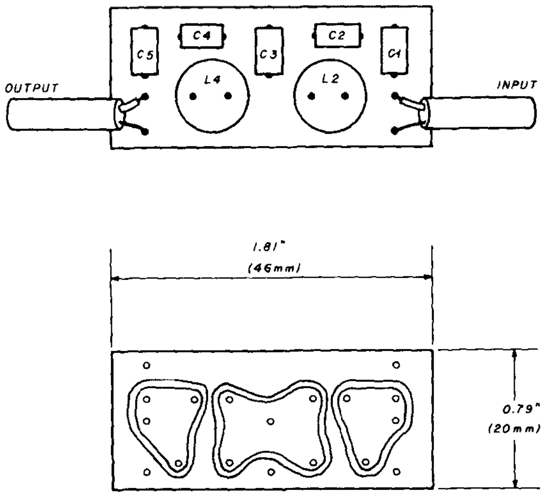


fig. 6. Stopband response of 1-dB/50-dB 5-branch 3-kHz elliptic low-pass filter, showing simulation and practical results.

transition band and stopband. I have used the term "transition band" to describe the response between the ripple cutoff frequency (f_{A_p}) and the start of stopband frequency (f_{A_s}). Many authors include this region in the stopband, but it was convenient for me to divide the total response into three regions, and the middle region seems most appropriately to be described as the transition band.

Curve *a* on **figs. 4, 5,** and **6** is the simulated response of the filter with theoretical values and ideal inductors (**table 1**, column 2). The response meets the 1-dB passband ripple (A_p), 3-kHz ripple cutoff frequency (f_{A_s}), 4221-Hz start of stopband frequency (f_{A_s}) and 50-dB minimum stopband attenuation (A_s) specifications. The calculations carried out to scale the 1-ohm, 1-rad/sec values (**table 1**, column 1) to 500 ohms, 3 kHz (**table 1**, column 2) are therefore correct. The frequencies of infinite attenuation in the stopband, f2 and f4, are 6480 Hz and 4378 Hz. These are the resonant frequencies of the parallel tuned circuits L2/C2 and L4/C4 respectively.

Curve *b* in **figs. 4, 5,** and **6** is the simulated response of the filter with rounded values (**table 1**, column 3). The passband ripple has increased, but by less than



Components for 1-dB/50-dB five-branch elliptic low-pass filter:

C1	0.22 μF Siemens metalized polyester 5 percent
C2	0.022 μF Siemens metalized polyester 5 percent
C3	0.27 μF Siemens metalized polyester 5 percent
C4	0.068 μF Siemens metalized polyester 5 percent
C5	0.18 μF Siemens metalized polyester 5 percent
L2	27 mH Toko 10RB
L4	18 mH Toko 10RB

fig. 7. PCB foil pattern and component placement for 5-branch elliptic low-pass filter.

table 1. Component values of 1-dB/50-dB five-branch elliptic low-pass filter. 500 ohm, 3-kHz values obtained by multiplying 1 ohm, 1 rad/sec values by 1.061×10^{-7} for capacitors and 2.653×10^{-2} for inductors.

component	column 1 1 ohm, 1 rad/sec value	column 2 500 ohm, 3 kHz theoretical value, ideal inductors	column 3 500 ohm, 3 kHz rounded value, ideal inductors	column 4 500 ohm, 3 kHz rounded value, real inductors
C1	1.933 F	0.2051 μF	0.22 μF	0.22 μF
C2	0.223 F	0.0237 μF	0.022 μF	0.022 μF
C3	2.392 F	0.2538 μF	0.27 μF	0.27 μF
C4	0.626 F	0.0664 μF	0.068 μF	0.068 μF
C5	1.635 F	0.1735 μF	0.18 μF	0.18 μF
L2	0.963 H	25.5 mH	27 mH	27 mH
L4	0.750 H	19.9 mH	18 mH	18 mH
R2	—	0 ohms	0 ohms	22 ohms
R4	—	0 ohms	0 ohms	17 ohms

0.5 dB; there has also been a slight decrease in the ripple cutoff frequency. In the transition band, close agreement with the unrounded value response is found. The frequency at which the stopband minimum attenuation is first achieved (f_{A_s}) is now 4330 Hz rather than the original 4221 Hz. The frequencies of infinite attenuation are now 6530 Hz and 4550 Hz, reflecting the change in resonant frequencies of L2/C2 and L4/C4 because of their value changes. One interesting result of the component changes is that the minimum stopband attenuation is now approximately 53 dB, rather than 50 dB. Though not shown in fig. 6, this value is maintained beyond 10 kHz. This increase in stopband attenuation seems to be a result of the increase in passband ripple, since these two parameters can be traded off at the design stage.

Curve *c* in figs. 4, 5, and 6 is the simulated response of the filter with rounded values and real inductors (table 1, column 4). The filter now exhibits non-zero insertion loss at all frequencies (having a value of about 0.3 dB at 100 Hz) and increased passband ripple. Other effects of the low-*Q* inductors are the smoothing of the final passband ripple and a gradual, rather than a sharp, transition from the passband into the transition band. The frequencies of high attenuation in the stopband are the same as those of the rounded values, ideal inductors' response (curve *b*), but the attenuations are no longer infinite. They are approximately 63 dB at 4550 Hz and 78 dB at 6530 Hz. The minimum stopband attenuation is similar to that shown by curve *b*, being approximately 53 dB.

Curve *d* in figs. 4, 5, and 6 shows the measured response obtained from the real filter constructed with the values of table 1, column 4. These results were obtained using a Hewlett Packard 3585A Spectrum Analyzer, which enables filter response measurements to be made very easily. Curve *d* shows close agreement with curve *c*, indicating that the simulation

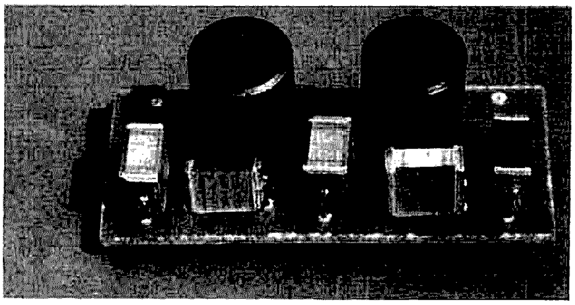


fig. 8. Experimental five-branch elliptic low-pass filter.

method was largely valid. The passband response is slightly better than simulations indicated, probably because the inductors have a higher *Q* than the simple simulation model produces. (Remember, this was seen from the measured *Q* values plotted in fig. 2.) Differences in the transition band and stopband response are probably because of the real components having values slightly different from their nominal values. The value of high attenuation attained by the practical filter at 4670 Hz is similar to that of the simulated filter (approximately 63 dB), but is approximately 5 dB less than the simulated filter at 6530 Hz. Minimum attenuation in the stopband is still a healthy 53 dB. This attenuation has been measured beyond 10 kHz, and remains at greater than 50 dB to more than 30 MHz.

Before I comment on the significance of the changes in performance between the theoretical and the final, practical filter, I would like to show the method used to construct this filter and the results obtained from a more complex low-pass design.

Figure 7 shows the PCB foil pattern and component layout used for this practical filter. The layout follows the schematic of the filter, with the output as far from the input as possible. A good ground plane

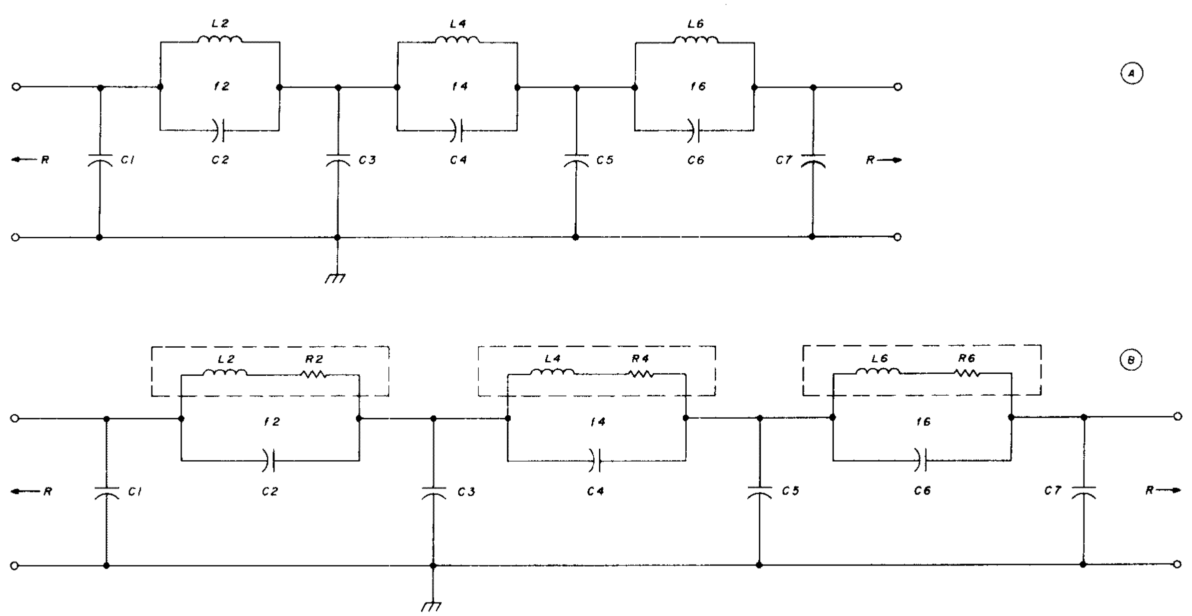


fig. 9. Elliptic 7-branch low-pass filter. (*A*) schematic of filter with ideal inductors, (*B*) schematic of filter with real inductors.

is maintained by etching away only the minimum amount of copper possible. The small amount of extra capacitance between each circuit node and ground is insignificant when compared with the capacitors. The ground plane gives low-impedance ground return paths and helps isolate the output from the input. If required, the filter can be made even more compact than my layout indicates. The ferrite caps over the Toko coils are such that they can be positioned very close to each other without any coupling effects. Stripboard can also be used for construction without any drastic effects on performance. **Figure 8** shows a photograph of this filter.

Figure 9 shows the schematics for a seven-branch elliptic low-pass filter, with and without the resistors to simulate the real inductors. The specification for this filter is:

passband ripple (A_p)	0.18 dB
stopband minimum attenuation (A_s)	50.1 dB
ripple cutoff frequency (f_{A_p})	3000 Hz
start of stopband frequency (f_{A_s})	3500 Hz
source impedance	500 ohms
load impedance	500 ohms

Table 2 shows the component values for the 1-ohm, 1-rad/sec prototype;[17] the exact values scaled to 500 ohms, 3 kHz; and the rounded values both with and without resistors to simulate the lossy real inductors. Again, each component has been rounded to the nearest E12 preferred value. Note that in the case of C4 and L4, both have been rounded downward, so we

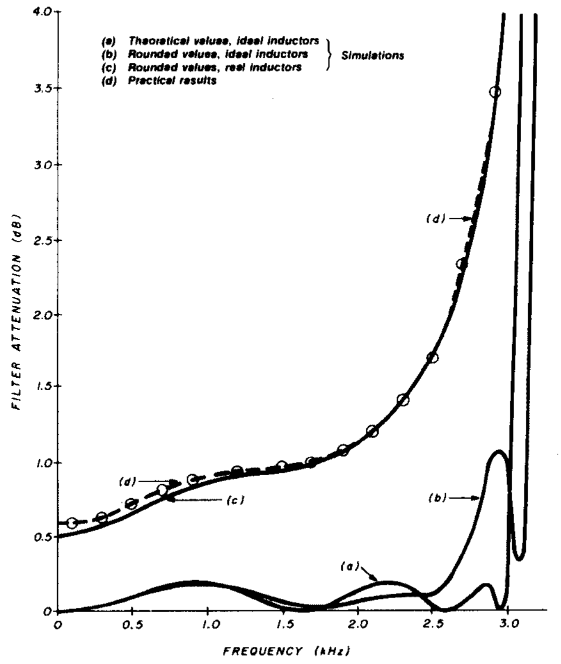


fig. 10. Passband response of 0.18-dB/50.1-dB 7-branch 3-kHz elliptic low-pass filter, showing simulation and practical results.

should expect a larger than average shift in the f4 frequency. If the peaks in stopband attenuation are being used in a particular application to obtain more than

the minimum stopband attenuation at particular frequencies, then it is best to round the tuned circuit elements in the filter in opposite directions. The frequencies of these peaks can be placed accurately by rounding only the inductor values and then selecting or paralleling capacitors to obtain the desired resonances.

The simulation and practical results for this filter are shown in **figs. 10, 11,** and **12.** I will not describe the responses in as much detail as for the 1-dB/50-dB low-

table 2. Component values of 0.18-dB/50.1-dB seven-branch elliptic low-pass filter. 500 ohm, 3-kHz values obtained by multiplying 1 ohm, 1 rad/sec values by 1.061 × 10^{-7} for capacitors and 2.653 × 10^{-2} for inductors.

component	column 1 1 ohm, 1 rad/sec value	column 2 500 ohm, 3 kHz theoretical value, ideal inductors	column 3 500 ohm, 3 kHz rounded value, ideal inductors	column 4 500 ohm, 3 kHz rounded value, real inductors
C1	1.183 F	0.1255 μF	0.12 μF	0.12 μF
C2	0.1853 F	0.0197 μF	0.018 μF	0.018 μF
C3	1.535 F	0.1629 μF	0.15 μF	0.15 μF
C4	0.9576 F	0.1016 μF	0.1 μF	0.1 μF
C5	1.307 F	0.1387 μF	0.15 μF	0.15 μF
C6	0.6755 F	0.0717 μF	0.068 μF	0.068 μF
C7	0.8543 F	0.0906 μF	0.082 μF	0.082 μF
L2	1.203 H	31.92 mH	33 mH	33 mH
L4	0.7482 H	19.85 mH	18 mH	18 mH
L6	0.8217 H	21.80 mH	22 mH	22 mH
R2	—	0 ohms	0 ohms	26 ohms
R4	—	0 ohms	0 ohms	17 ohms
R6	—	0 ohms	0 ohms	19 ohms

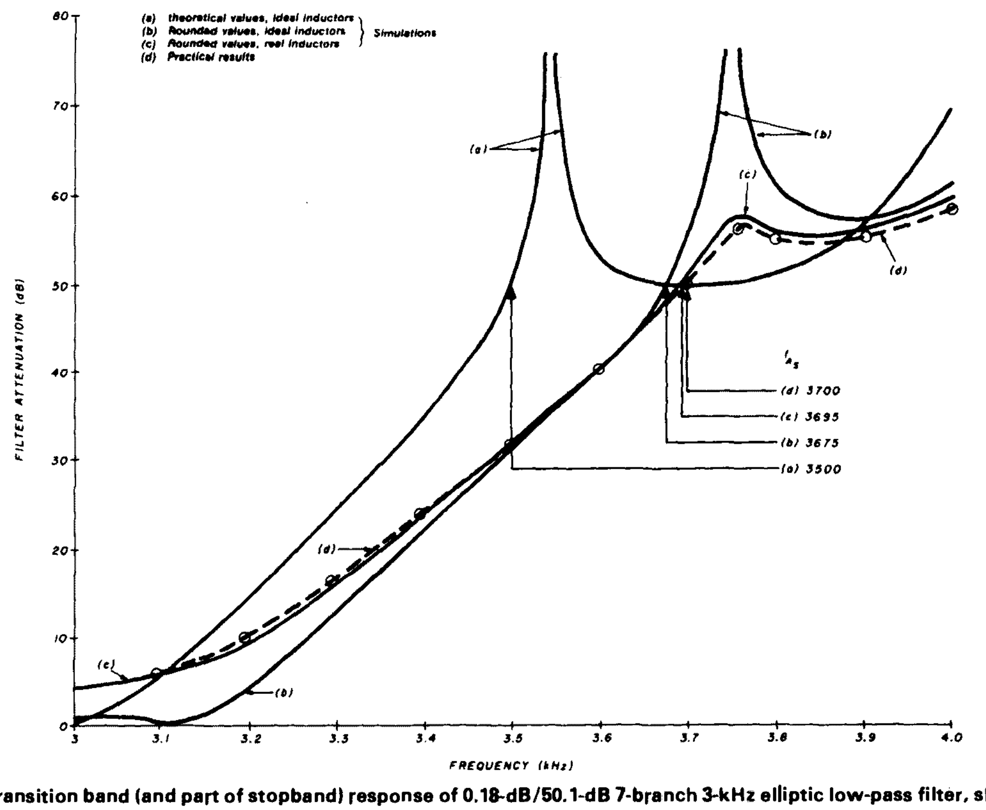


fig. 11. Transition band (and part of stopband) response of 0.18-dB/50.1-dB 7-branch 3-kHz elliptic low-pass filter, showing simulation and practical results.

pass filter; the graphs speak for themselves. The scaled, theoretical value response meets the filter specification exactly. As before, the rounded values, ideal inductor response is very similar to that of the the theoretical values apart from an increase in passband ripple (especially close to the cutoff frequency), a shift in the f2 frequency from 3500 Hz to 3751 Hz, and a corresponding shift in the initial 50.1-dB attenuation frequency (f_{A_s}). The minimum stopband attenuation is approximately 52 dB.

table 3. Component values of 0.18-dB/81-dB seven-branch elliptic low-pass filter. 500-ohm, 3-kHz values obtained by multiplying 1 ohm, 1 rad/sec values by 1.061 × 10^{-7} for capacitors and 2.653 × 10^{-2} for inductors.

component	column 1 1 ohm 1 rad/sec value	column 2 500 ohm, 3 kHz theoretical value, ideal inductors	column 3 500 ohm, 3 kHz rounded value, real inductors
C1	1.280 F	0.1358 μF	0.15 μF
C2	0.065 F	0.0690 μF	0.0068 μF
C3	1.943 F	0.2062 μF	0.22 μF
C4	0.3079 F	0.0327 μF	0.033 μF
C5	1.837 F	0.1949 μF	0.18 μF
C6	0.2183 F	0.0232 μF	0.022 μF
C7	1.145 F	0.1215 μF	0.12 μF
L2	1.321 H	35.05 mH	33 mH
L4	1.183 H	31.38 mH	33 mH
L6	1.157 H	30.70 mH	33 mH
R2	—	0 ohms	26 ohms
R4	—	0 ohms	26 ohms
R6	—	0 ohms	26 ohms

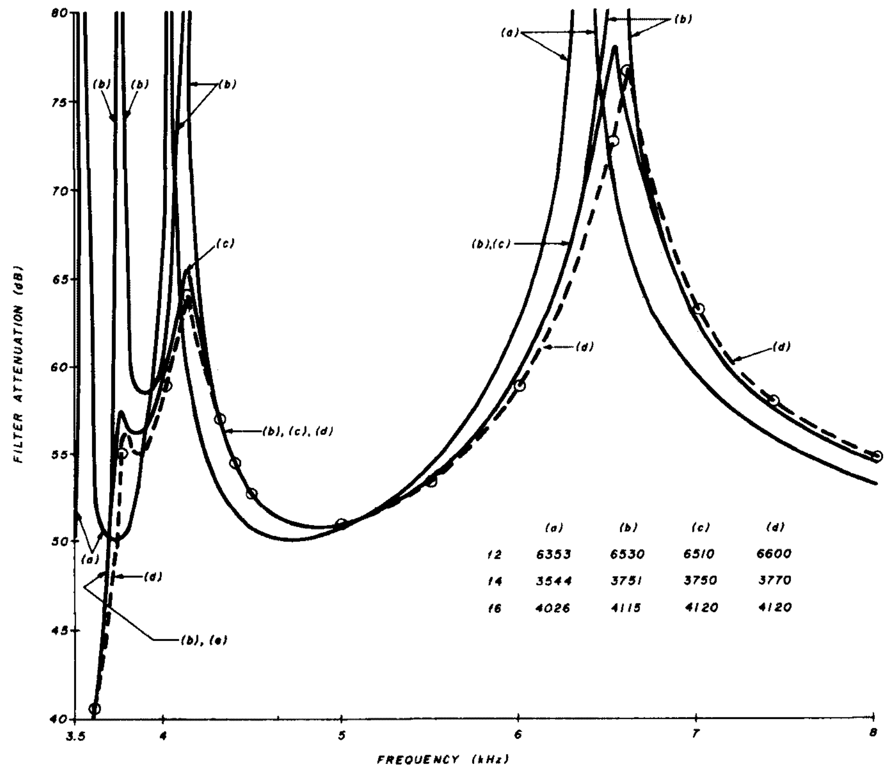
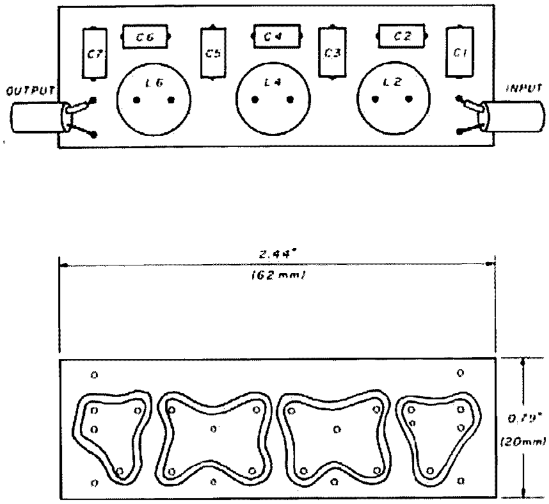


fig. 12. Stopband response of 0.18-dB/50.1-dB 7-branch 3-kHz elliptic low-pass filter, showing simulation and practical results.



Components for 0.18-dB/50 1-dB seven-branch elliptic low-pass filter

C1 0.12 μF Siemens metalized polyester 5 percent
C2 0.018 μF Siemens metalized polyester 5 percent
C3 0.15 μF Siemens metalized polyester 5 percent
C4 0.1 μF Siemens metalized polyester 5 percent
C5 0.15 μF Siemens metalized polyester 5 percent
C6 0.068 μF Siemens metalized polyester 5 percent
C7 0.082 μF Siemens metalized polyester 5 percent
L2 33 mH Toko 10RB
L4 18 mH Toko 10RB
L6 22 mH Toko 10RB

fig. 13. PCB for foil pattern and component placement for 7-branch elliptic low-pass filtar.

There is close agreement between the simulated rounded values, real inductors' filter response, curve *c*, and the practical results, curve *d*, indicating accurate modeling of the circuit. Insertion loss at 100 Hz is approximately 0.6 dB. Rounding of the passband edge because of the low-Q inductors can clearly be seen. The maximum in stopband attenuation at approximately 3760 Hz is now barely perceptible. Minimum stopband attenuation is approximately 52 dB (compared to the theoretical value of 50.1 dB). With this filter, the stopband attenuation reaches a minimum at a frequency above 10 kHz then increases again to more than 70 dB at 30 MHz.

Figure 13 shows the PCB foil pattern and component layout used for this filter, and **fig. 14** is a photograph of my prototype. Again, if required, this filter can be made considerably smaller.

high stopband attenuation low-pass filters

The two low-pass filters considered so far were specified to have a minimum stopband attenuation of 50 dB and achieved that figure. In theory, at the points of maximum attenuation in the stopband, these filters should have infinite attenuation, but the finite Q of the inductors causes this attenuation to be reduced to typically 55-75 dB, depending on which attenuation peak is considered. Can, therefore, filters with high stopband attenuation be constructed with these inductors?

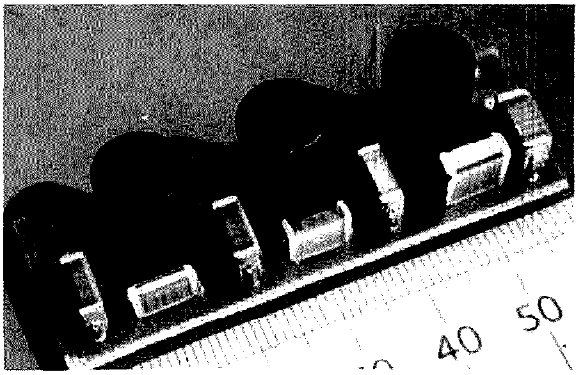


fig. 14. Experiinental seven-branob elliptic low-pass filter.

To answer this question, a 0.18-dB passhand ripple, 81-dB minimum stopband attenuation seven-branch elliptic low-pass filter was selected[18] and scaled to 500 ohms, 3 kHz. **Table 3** shows the component values for this filter. The expected start of stopband frequency (f_{A_S}) is 4872 Hz, corresponding to a

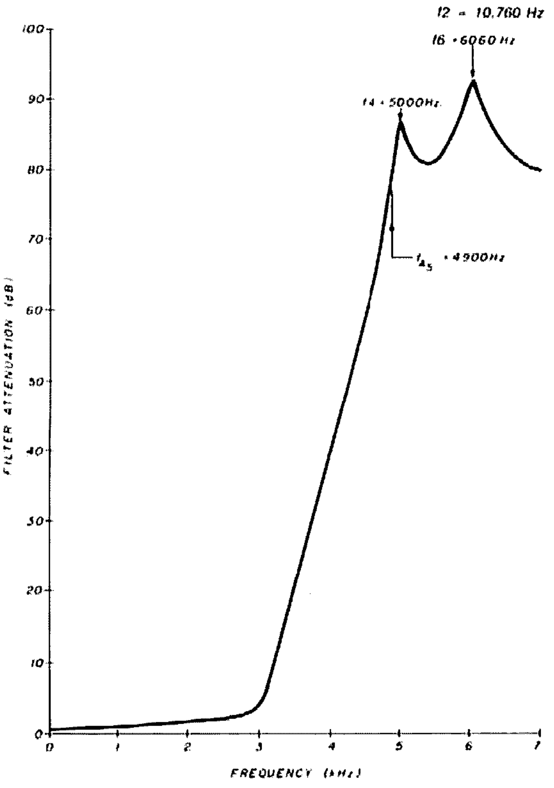


fig. 15. 0-7.5 kHz responsa of 0.18-dB/81-dB 7-branch elliptic low-pass filter sbowing practical iesults only.

normalized start of stopband *angular* frequency of 1.624 radians/second.

Detailed simulation results for this filter are not given here and I have only shown the practical results when the layout of **fig. 13** was used. These results are shown in **fig. 15.**

As we expected, the actual passband response has greater ripple (exceeds the 0.18 dB theoretical value) and has rounded corners. The insertion loss at 100 Hz is approximately 0.4 dB. Stopband attenuation shows a minimum of approximately 80 dB above the measured (f_{A_s}) of 4900 Hz. The two maxima of stopband attenuation below 10 kHz occur at approximately 5000 Hz and 6060 Hz, with values of 86 and 92 dB respectively. A further maximum of 93 dB occurs at 10,760 Hz.

This example shows that despite the use of low-*Q* inductors, high stopband attenuation can still be obtained if required in a particular application.

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appendix
elliptic low-pass filter terminology

The schematic of a five-branch elliptic low-pass filter is shown in **fig. A1(A)**. The ideal response of this filter is shown in **fig. A1 (B)**.

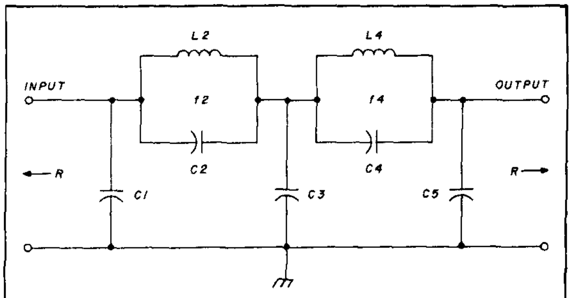

fig. A1(A). Schematic diagram of a 5-branch elliptic low-pass filter.

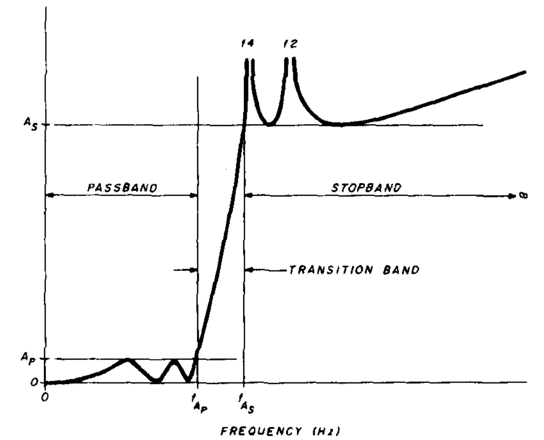

fig. A1(B). Ideal response of a 5-branch elliptic low-pass filter.

Some of the terms used in describing elliptic filters are explained below.

- **Passband ripple.** A variation in the passband attenuation between zero and some maximum value, A_p. The number of half cycles of ripple is equal to the number of branches of the filter.
- **Minimum stopband attenuation.** The value of stopband attenuation, designated A_s, below which the attenuation never falls having once achieved this value.
- **Maxima in stopband attenuation.** Frequencies within the stopband region, designated f2 and f4, at which the attenuation is infinite due to the resonances of L2/C2 and L4/C4 respectively.
- **Ripple cutoff frequency.** The frequency, designated f_{A_p}, at which the attenuation first exceeds the A_p value.
- **Start of stopband frequency.** The frequency, designated f_{A_s}, at which the attenuation first exceeds the A_s value.

Tables of component values for filter designs usually give normalized values; that is, values which produce a filter with source and load impedance values of 1 ohm and a cutoff frequency (angular) of 1 rad/sec. To obtain practical component values for impedance of R ohms and a cutoff frequency of f Hz, *multiply all inductor values by* $\frac{R}{2\pi f}$ and all capacitor values by $\frac{1}{2\pi fR}$.

ham radio

adjusting SSB amplifiers

Use a tuning pulser for added safety and linearity

How many times have you had your on-the-air conversations interrupted by someone who's more than 5 kHz from the frequency you're using? I can't remember how many times this has happened to me.

Splatter is rude and unnecessary. It *ought* to be against FCC rules — but it isn't. Part 97 of the Amateur Radio Rules states, in 97.73, (under "purity of emissions," paragraph *d*) that spurious radiation causing harmful interference to the reception of another radio station is against the rules, and that the licensee responsible for it may be required to eliminate the interference. At first this sounds good, but there's a note that explains that they're only talking about the garbage that slips outside of the Amateur band. In other words, if an operator on 14,275 kHz splatters ± 50 kHz, he's in compliance with Part 97 because his spurious emissions *inside* the Amateur Radio bands are specifically exempted by the note that follows paragraph *d*. If the same splattering station were to move up to 14,310 kHz, the splatter would extend beyond the Amateur band and a violation of the rules would occur.

It seems, then, that the only way to avoid stations who splatter would be to operate near the band edge with a clean signal. Dirty stations operating in this area are likely to spill over the band edge and receive FCC citations.

The most common cause of splatter is a mis-adjusted "linear" amplifier. Linear amplifiers are far from foolproof; in fact, they're easy to operate non-linearly. There are many different settings of the tune and load controls, and only one combination will represent the best compromise between linearity and power output.

Without a doubt, the two-tone test, used in conjunction with an oscilloscope, is the best method of adjusting a linear amplifier. Unfortunately, the two-tone generator and a good oscilloscope cost more than many linear amplifiers.

A second method is the single-tone test. This involves inserting a single frequency — usually with the CW mode on the transmitter, or transceiver — at the maximum carrier level, and adjusting the linear amplifier for maximum power output by alternately adjusting the plate-tune and load controls. This method can be improved if the loading control is increased (for less capacitance) slightly beyond the point where the power peaks. At this point the linearity will be improved and the power output drops, just perceptibly. But this method places a terrible strain on the amplifier tubes and power supply. If the operator makes an error during this procedure, the tubes are likely to be damaged because the duty cycle is 100 percent.

Another problem with the single-tone test is that the average amplifier power supply is not designed to supply the 2500 watts of key-down power needed for the legal output of 1500 watts, assuming a typical efficiency of 60 percent. The average table-top 1500 watt output linear amplifier appears to have a 600 watt CCS plate transformer. This is completely satisfactory in SSB service, but it is not intended for full throttle single-tone tune-up.

Reducing the carrier level on tune-up would reduce the strain, but reduced carrier level results in reduced peak plate current in the amplifier tubes. Because peak plate current and peak AC plate voltage determine the plate load impedance, altering one would upset the

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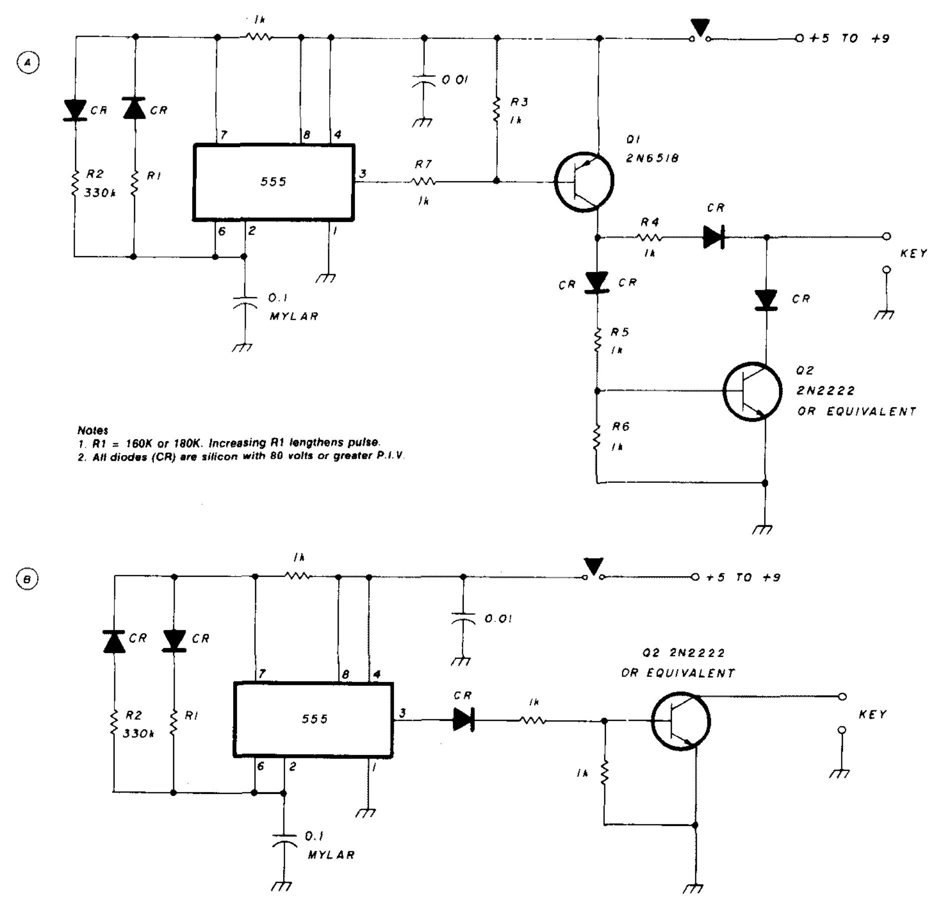

fig. 1. Schematic of a tuning pulser circuit. (A) for positive or negative keyed transmitters, and (B) for positive-keyed transmitters only.

adjustment, since the plate tune and load controls would be set to match a different plate load than will be encountered on voice peaks. This mis-adjustment method will make the amplifier splatter and beat up the grids of the common grounded-grid, class AB_2 amplifier because of excessive grid current.

Yet another problem with the single tone, or A-zero (AØ) tune-up is that it's not permitted, except for short periods below 51 MHz — but the chance of being cited for AØ emission is probably slim in comparison to the chance of experiencing a melt-down.

So the problem is how to tune-up a linear amplifier, linearly and safely, without spending a lot of money. The solution is obviously some device that will produce maximum peak plate current with a human voice duty-cycle, which is about 33 percent on the average. For ease of adjustment, the device must also produce steady meter indications.

The circuit shown in **fig. 1A** is a solution to the problem. It consists of a 555 timer and two transistor switches, one for positive keyed transmitters and the other for negative keyed transmitters. The 555 timing resistors have steering diodes so that the charge and discharge times are different by the ratio of 2:1. This produces a duty cycle of 0.3333, or one-third. The output of the 555 is high for about 24 ms and low for about 12 ms. When the base of Q1 goes low, it is forward biased and conducts collector current through R5 and the base of Q2, whose collector is then switched on. During the 24 ms high interval, the base-to-emitter voltage of Q1 is zero, which turns Q1 off, which then turns Q2 off. This causes the amplifier to work hard for 12 ms, then rest for 24 ms, alternately working and resting until the tuning process is complete.

The circuit shown in **fig. 1B**, which will key positive voltages only, requires fewer parts than the previously described circuit. (Most transceivers built today are positive-keyed.)

The pulser should be enclosed in a metal box or open-button metal chassis. (I used a 3 × 5 × 1 inch/7

× 12 × 2.5 cm open bottom chassis.) The keying lead to the transceiver should be shielded. These precautions will help to keep RF, which would interfere with proper operation, out of the pulser.

A good method of powering the pulser is to use four AA penlight batteries. There's plenty of room in the box for a four-cell AA holder, and the added milliampere hour capacity advantage over a 9-volt battery makes the AA cells a good choice.

Either circuit can be built on a piece of perfboard in less than one hour, so it isn't worth the trouble of making a printed circuit board unless you're going to make several tuning pulsers for Christmas presents.

using the tuning pulser

Set the transceiver to CW. Turn on the pulser and advance the carrier control until you have some ALC. If your exciter has an individual band tune feature, adjust it for maxiumum relative output.

Now turn on your amplifier and set the plate voltage to SSB. Key the pulser again and tune for maximum relative output on the plate tune and load controls. Sometimes a slight improvement in linearity can be had by loading the amplifier past the peak in output until the relative output drops about 2 percent.

Normal readings of plate current using the tuning pulser should be approximately one-third of what you would see with a steady carrier. Average-reading RF wattmeters are good for relative indications only, when tuning with the pulser. The indicated wattage cannot be simply multiplied by 3 to find the peak power because the time constant of the meter is unknown.

If you own a linear amplifier that uses sweep tubes, which are easily damaged by normal tune-up methods, a tuning pulser could provide increased tube life.

Another use for the tuning pulser is in locating loose hardware (such as clotheslines, rain gutters, or rusty guy-wires), that may be causing TVI by metal-oxide diode-rectification. Such problems usually occur only at or near the peak-power level. Running the typical amplifier key-down for the length of time needed to locate the fault is risky. With the tuning pulser, the amplifier can generate full peak power with little risk of damage, even for lengthy periods of time. A harmonic "sniffing" device is then used to determine the location of the fault. (Normal operating courtesies still apply; check the frequency first with your receiver, keep the tests short, and try to make your tests when the band is dead. — Ed)

ham radio

understanding telephones

Everybody has one. But what makes it work?

Although telephones and telephone company practices may vary dramatically from one country to another, and even from one locality to another, the basic principles underlying the way they work remain unchanged.

Every telephone consists of three separate subassemblies, each capable of independent operation. These assemblies are the *speech network*, the *dialing mechanism*, and the *ringer* or *bell*. Together, these parts — as well as any additional devices such as modems, dialers, and answering machines — are attached to the *phone line*.

the phone line

A telephone is usually connected to the telephone exchange by about three miles (4.83 km) of a twisted pair of No. 22 (AWG) or 0.5 mm copper wires, known by your phone company as "the loop." Although copper is a good conductor, it does have resistance. The resistance of No. 22 AWG wire is 16.46 ohms per thousand feet at 77 degrees F (25 degrees C). In the United States, wire resistance is measured in ohms per thousand feet; telephone companies describe loop length in kilofeet (thousands of feet). In other parts of the world, wire resistance is usually expressed as Ohms per kilometer.

Because telephone apparatus is generally considered to be current driven, all phone measurements refer to current *consumption*, not voltage. The length of the wire connecting the subscriber to the telephone exchange affects the total amount of current that can be drawn by anything attached at the subscriber's end of the line.

In the United States, the voltage applied to the line to drive the telephone is 48 VDC; some countries use 50 VDC. Note that telephones are peculiar in that the signal line is also the power supply line. The voltage is supplied by lead acid cells, thus assuring a hum-free supply and complete independence from the electric company, which may be especially useful during power outages.

At the telephone exchange the DC voltage and audio signal are separated by directing the audio signal through 2 μF capacitors and blocking the audio from the power supply with a 5-Henry choke in each line. Usually these two chokes are the coil windings of a relay that switches your phone line at the exchange; in the United States, this relay is known as the "A" relay (see **fig. 1**). The resistance of each of these chokes is 200 ohms.

We can find out how well a phone line is operating by using Ohm's law and an ammeter. The DC resistance of any device attached to the phone line is often quoted in telephone company specifications as either 600 or 900 ohms; this will vary from one country to another. But the resistance may, in fact, vary from 300 to 1,000 ohms.

Using these figures you can estimate the distance between your telephone and the telephone exchange. In the United States, the telephone company guarantees you no lower current than 20 mA — or what is known to your phone company as a "long loop." A "short loop" will draw 50 to 70 mA, and an average

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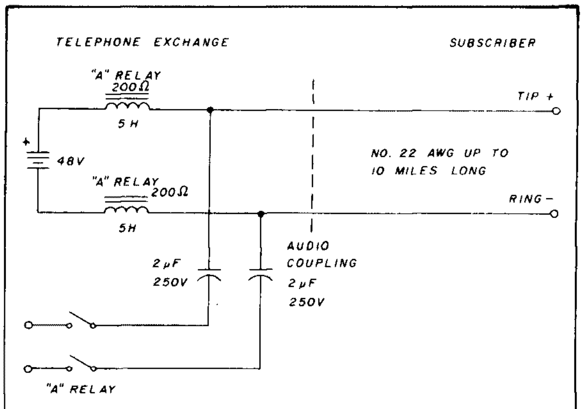


fig. 1. The phone line.

loop, about 35 mA. Some countries will consider their maximum loop as low as 12 mA. In practice, United States telephones are usually capable of working at currents as low as 14 mA. Some exchanges will consider your phone in use and feed dial tone down the line with currents as low as 8 mA, even though the telephone may not be able to operate.

Although the telephone company has supplied plenty of nice clean DC direct to your home, don't assume you have a free battery for your own circuits. The telephone company wants the DC resistance of your line to be about 10 Megohms when there's no apparatus in use ("on hook," in telephone company jargon); you can draw no more than 5 microamperes while the phone is in that state. When the phone is in use, or "off hook," you can draw current, but you'll need that current to power your phone. Any current you might draw for other purposes would tend to lower the signal level.

The phone line has an impedance composed of distributed resistance, capacitance, and inductance. The impedance will vary according to the length of the loop, the type of insulation of the wire, and whether the wire is aerial cable, buried cable, or bare parallel wires strung on telephone poles. For calculation and specification purposes, the impedance is normally assumed to be 600 or 900 ohms. If the instrument attached to the phone line should be of the wrong impedance, you'd get a mismatch, or what telephone company personnel refer to as "return loss." (Radio Amateurs will recognize return loss as SWR.) A mismatch on telephone lines results in echo and whistling, which the phone company calls "singing" and owners of very cheap telephones may have come to expect. A mismatched device can, by the way, be matched to the phone line by placing resistors in parallel or series with the line to bring the impedance of the device to within the desired limits. This will cause some signal loss, of course, but will make the device usable.

A phone line is a balanced feed, with each side equally balanced to ground. Any imbalance will introduce hum and noise to the phone line and increase susceptability to RFI.

The balance of the phone line is known to your telephone company as "longitudinal balance." If both impedance match and balance to ground are kept in mind, any device attached to the phone line will perform well, just as the correct matching of transmission lines and devices will ensure good performance in radio practice.

If you live in the United States, the two phone wires connected to your telephone should be red and green. (In other parts of the world they may be different colors.) The red wire is negative and the green wire is positive. Your telephone company calls the green wire "Tip" and the red wire "Ring." (In other parts of the world, these wires may be called "A" and "B".) Most installations have another pair of wires, yellow and black. These wires can be used for many different purposes, if they're used at all. Some party lines use the yellow wire as a ground; sometimes there's 6.8 VAC on this pair to light the dials of Princess® type phones. If you have two separate phone lines (not extensions) in your home, you'll find the yellow and black pair carrying a second telephone line. In this case, black is "Tip" and yellow is "Ring."

The above description applies to a standard line with a DC connection between your end of the line and the telephone exchange. Most phone lines in the world are of this type, known as a "metallic line." In a metallic line, there may or may not be inductance devices placed in the line to alter the frequency response of the line; the devices used to do this are called "loading coils." (Note: if they impair the operation of your modem, your telephone company can remove them.) Other types of lines are party lines, which may be metallic lines but require special telephones to allow the telephone company to differentiate between subscribers. Very long lines may have amplifiers, sometimes called "loop extenders"on them. Some telephone companies use a system called "subscriber carrier," which is basically an RF system in which your telephone signal is heterodyned up to around 100 kHz and then sent along another subscriber's "twisted pair."

If you have questions about your telephone line, you can call your telephone company; depending on the company and who you can reach, you may be able to obtain a wealth of information.

the speech network

The speech network — also known as the "hybrid" or the "two wire/four wire network" — takes the incoming signal and feeds it to the earpiece and takes

the microphone output and feeds it down the line. The standard network used all over the world is an LC device with a carbon microphone; some newer phones use discrete transistors or ICs.

One of the advantages of an LC network is that it has no semiconductors, is not voltage sensitive, and will work continuously as the voltage across the line is reduced. Many transistorized phones stop working as the voltage approaches 3 to 4 volts.

When a telephone is taken off the hook, the line voltage drops from 48 volts to between 9 and 3 volts, depending on the length of the loop. If another telephone in parallel is taken off the hook, the current consumption of the line will remain the same and the voltage across the terminals of both telephones will drop. Bell Telephone specifications state that three telephones should work in parallel on a 20 mA loop; transistorized phones tend not to pass this test, although some manufacturers use ICs that will pass. Although some European telephone companies claim that phones working in parallel is "technically impossible," and discourage attempts to make them work that way, some of their telephones *will* work in parallel.

While low levels of audio may be difficult to hear, overly loud audio can be painful. Consequently, a well designed telephone will automatically adjust its transmit and receive levels to allow for the attenuation — or lack of it — caused by the length of the loop. This adjustment is called "loop compensation." In the United States, telephone manufacturers achieve this compensation with silicon carbide varistors that consume any excess current from a short loop (see **fig. 2**). Although some telephones using ICs have built-in loop compensation, many do not; the latter have been designed to provide adequate volume on the average loop, which means that they provide low volume on long loops, and are too loud on short loops. Various countries have different specifications for transmit and receive levels; some European countries require a higher transmit level than is standard in the United States so a domestically-manufactured telephone may suffer from low transmit level if used on European lines without modification.

Because a telephone is a duplex device, both transmitting and receiving on the same pair of wires, the speech network must ensure that not too much of the caller's voice is fed back into his or her receiver. This function, called "sidetone," is achieved by phasing the signal so that some cancellation occurs in the speech network before the signal is fed to the receiver. Callers faced with no sidetone at all will consider the phone "dead." Too little sidetone will convince callers that they're not being heard and cause them to shout, "I can hear you. Can you hear *ME?*" Too much sidetone causes callers to lower their voices and not be heard well at the other end of the line.

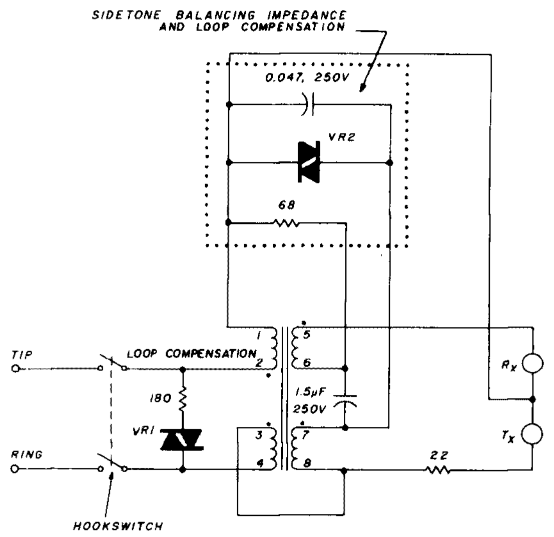


fig. 2. Telephone speech network. (Note: while *ham radio* designates varistors as *CR*, telephone companies customarily use the designation *VR*. — Ed.)

A telephone on a short loop with no loop compensation will appear to have too much sidetone, and callers will lower their voices. In this case, the percentage of sidetone is the same, but as the overall level is higher the sidetone level will also be higher.

the dial

There are two types of dials in use around the world. The most common one is called pulse, loop disconnect, or rotary; the oldest form of dialing, it's been with us since the 1920's. The other dialing method, more modern and much loved by Radio Amateurs is called Touchtone Dual Tone Multi-Frequency (DTMF) or Multi-Frequency (MF).

Pulse dialing is traditionally accomplished with a rotary dial, which is a speed governed wheel with a cam that opens and closes a switch in series with your phone and the line. It works by actually disconnecting or "hanging up" the telephone at specific intervals. The United States standard is one disconnect per digit, so if you dial a "1," your telephone is "disconnected" once. Dial a seven and you'll be "disconnected" seven times; dial a zero, and you'll "hang up" ten times. Some countries invert the system so "1" causes ten "disconnects" and 0, one disconnect. Some add a digit so that dialing a 5 would cause six disconnects and 0, eleven disconnects. There are even some systems in which dialing 0 results in one disconnect, and all other digits are *plus one*, making a 5 cause six disconnects and 9, ten disconnects.

Although most exchanges are quite happy with rates of 6 to 15 Pulses Per Second (PPS), the phone company accepted standard is 8 to 10 PPS. Some

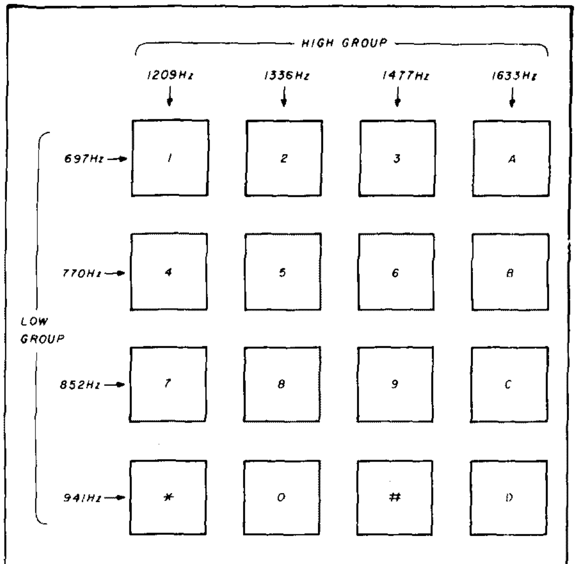


fig. 3. Standard DTMF pad and frequencies.

modern digital exchanges, free of the mechanical inertia problems of older systems, will accept a PPS rate as high as 20.

Besides the PPS rate, the dialing pulses have a make/break ratio, usually described as a percentage, but sometimes as a straight ratio. The North American standard is 60/40 percent; most of Europe accepts a standard of 63/37 percent. This is the pulse measured at the telephone, not at the exchange, where it's somewhat different, having travelled through the phone line with its distributed resistance, capacitance, and inductance. In practice, the make/break ratio does not seem to affect the performance of the dial when attached to a normal loop. Bear in mind that each pulse is a switch connect and disconnect across a complex impedance, so the switching transient often reaches 300 volts. Try not to have your fingers across the line when dialing.

Most pulse dialing phones produced today use a CMOS IC and a keyboard. Instead of pushing your finger round in circles, then removing your finger and waiting for the dial to return before dialing the next digit, you punch the button as fast as you want. The IC stores the number and pulses it out at the correct rate with the correct make/break ratio and the switching is done with a high-voltage switching transistor. Because the IC has already stored the dialed number in order to pulse it out at the correct rate, it's a simple matter for telephone designers to keep the memory "alive" and allow the telephone to store, recall, and redial the Last Number Dialed (LND). This feature enables you to redial by picking up the handset and pushing just one button.

Because pulse dialing entails rapid connection and disconnection of the phone line, you can "dial" a telephone that has lost its dial, by hitting the hookswitch rapidly. It requires some practice to do this with consistent success, but it can be done. A more sophisticated approach is to place a Morse key in series with the line, wire it as normally closed and send strings of dots corresponding to the digits you wish to dial.

Touchtone, the most modern form of dialing, is fast and less prone to error than pulse dialing. Compared to pulse, its major advantage is that its audio band signals can travel down phone lines further than pulse, which can travel only as far as your local exchange. Touchtone can therefore send signals around the world via the telephone lines, and can be used to control phone answering machines and computers. Pulse dialing is to touchtone as FSK or AFSK RTTY is to Switched Carrier RTTY, where mark and space are sent by the presence or absence of DC or unmodulated RF carrier. Most Radio Amateurs are familiar with DTMF for controlling repeaters and for accessing remote and auto phone patches.

Bell Labs developed DTMF in order to have a dialing system that could travel across microwave links and work rapidly with computer controlled exchanges. Each transmitted digit consists of two separate audio tones that are mixed together (see **fig. 3**). The four vertical columns on the keypad are known as the high group and the four *horizontal* rows as the low group; the digit 8 is composed of 1336 Hz and 852 Hz. The level of each tone is within 3 dB of the other, (the telephone company calls this "Twist"). A complete touchtone pad has 16 digits, as opposed to ten on a pulse dial. Besides the numerals 0 to 9, a DTMF "dial" has *, #, A, B, C, and D. Although the letters are not normally found on consumer telephones, the IC in the phone is capable of generating them.

The * sign is usually called "star" or "asterisk." The # sign, often referred to as the "pound sign." is actually called an *octothorpe*. Although many phone users have never used these digits — they're not, after all, ordinarily used in dialing phone numbers — they are used for control purposes, phone answering machines, bringing up remote bases, electronic banking, and repeater control. The one use of the octothorpe that may be familiar occurs in dialing international calls from phones in the United States. After dialing the complete number, dialing the octothorpe lets the exchange know you've finished dialing. It can now begin routing your call; without the octothorpe, it would wait and "time out" before switching your call.

When DTMF dials first came out they had complicated cams and switches for selecting the digits and used a transistor oscillator with an LC tuning network to generate the tones. Modern dials use a matrix switch and a CMOS IC that synthesizes the tones from

a 3.57 MHz (TV color burst) crystal. This oscillator runs only during dialing, so it doesn't normally produce QRM.

Standard DTMF dials will produce a tone as long as a key is depressed. No matter how long you press, the tone will be decoded as the appropriate digit. The shortest duration in which a digit can be sent and decoded is about 100 milliseconds (ms). It's pretty difficult to dial by hand at such a speed, but automatic dialers can do it. A twelve-digit long distance number can be dialed by an automatic dialer in a little more than a second — about as long as it takes a pulse dial to send a single 0 digit.

The output level of DTMF tones from your telephone should be between 0 and −12 dBm. In telephones, 0 dB is 1 milliwatt over 600 ohms. So 0 dB is 0.775 volts. Because your telephone is considered a 600 ohm load, placing a voltmeter across the line will enable you to measure the level of your tones.

the ringer

Simply speaking this is a device that alerts you to an incoming call. It may be a bell, light, or warbling tone. The telephone company sends a ringing signal which is an AC waveform. Although the common frequency used in the United States is 20 Hz, it can be any frequency between 15 and 68 Hz. Most of the world uses frequencies between 20 and 40 Hz. The voltage at the subscriber's end depends upon loop length and number of ringers attached to the line; it could be between 40 and 150 volts. Note that ringing voltage can be hazardous; when you're working on a phone line, be sure at least one telephone on the line is off the hook (in use); if any are not, take high voltage precautions. The telephone company may or may not remove the 48 VDC during ringing; as far as you're concerned, this is not important. *Don't take chances.*

The ringing cadence — the timing of ringing to pause — varies from company to company. In the United States the cadence is normally 2 seconds of ringing to 4 seconds of pause. An unanswered phone in the United States will keep ringing until the caller hangs up. But in some countries, the ringing will "time out" if the call is not answered.

The most common ringing device is the gong ringer, a solenoid coil with a clapper that strikes either a single or double bell. A gong ringer is the loudest signaling device that is solely phone-line powered.

Modern telephones tend to use warbling ringers, which are usually ICs powered by the rectified ringing signal. The audio transducer is either a piezoceramic disk or a small loudspeaker via a transformer.

Ringers are isolated from the DC of the phone line by a capacitor. Gong ringers in the United States use a 0.47 μF capacitor. Warbling ringers in the United States generally use a 1.0 μF capacitor. Telephone companies in other parts of the world use capacitors between 0.2 and 2.0 μF. The paper capacitors of the past have been replaced almost exclusively with capacitors made of Mylar film. Their voltage rating is always 250 volts.

The capacitor and ringer coil, or Zeners in a warbling ringer, constitute a resonant circuit. When your phone is hung up ("on hook") the ringer is across the line; if you've turned off the ringer you've merely silenced the transducer, not removed the circuit from the line.

When the telephone company uses the ringer to test the line, it sends a low-voltage, low-frequency signal down the line (usually 2 volts at 10 Hz) to test for continuity. The company keeps records of the expected signals on your line. This is how it can tell if you've added equipment to your line. If your telephone has had its ringer disconnected, the telephone company cannot detect its presence on the line.

Because there is only a certain amount of current available to drive ringers, if you keep adding ringers to your phone line you'll reach a point at which either all ringers will cease to ring, some will cease to ring, or some ringers will ring weakly. In the United States the phone company will guarantee to ring five normal ringers. A normal ringer is defined as a standard gong ringer as supplied in a phone company standard desk telephone. The value given to this ringer is *Ringer Equivalence Number (REN) 1.* If you look at the FCC registration label of your telephone, modem, or other device to be connected to the phone line, you'll see the *REN* number. It can be as high as 3.2, which means that device consumes the equivalent power of 3.2 standard ringers, or 0.0, which means it consumes no current when subjected to a ringing signal. If you have problems with ringing, total up your RENs; if the total is greater than 5, disconnect ringers until your REN is at 5 or below.

Other countries have various ways of expressing REN, and some systems will handle no more than three of their standard ringers. But whatever the system, if you add extra equipment and the phones stop ringing, or the phone answering machine won't pick up calls, the solution is disconnect ringers until the problem is resolved. Warbling ringers tend to draw less current than gong ringers, so changing from gong ringers to warbling ringers may help you spread the sound better.

Frequency response is the second criterion by which a ringer is described. In the United States most gong ringers are electromechanically resonant. They are usually resonant at 20 and 30 Hz (±3 Hz). The FCC refers to this as *A* so a normal gong ringer is described as *REN 1.0A*. The other common frequency response is known as type B. Type B ringers will respond to signals between 15.3 and 68.0 Hz. Warbling ringers are

all type B and some United States gong ringers are type B. Outside the United States, gong ringers appear to be non-frequency selective, or type B.

Because a ringer is supposed to respond to AC waveforms, it will tend to respond to transients (such as switching transients) when the phone is hung up, or when the rotary dial is used on an extension phone. This is called "bell tap" in the United States; in other countries, it's often called "bell tinkle." While European and Asian phones tend to bell tap, or tinkle, United States ringers that bell tap are considered defective. The bell tap is designed out of gong ringers and fine tuned with bias springs. Warbling ringers for use in the United States are designed not to respond to short transients; this is usually accomplished by rectifying the AC and filtering it before it powers the IC, then not switching on the output stage unless the voltage lasts long enough to charge a second capacitor.

conclusion

This brief primer describing the working parts of a telephone is intended to provide a better understanding of phone equipment. Note that most telephone regulatory agencies, including the FCC, forbid modification of anything that has been previously approved or attached to phone lines.

ham radio

a DTMF controller for repeaters

This multifunctional, expandable decoder uses high quality components — but doesn't cost a fortune to build

Remote control of the various functions of a repeater is usually accomplished through the use of dual-tone multifrequency (DTMF) signals generated in a pushbutton touchtone pad. These signals can also be generated in computer-controlled synthesizers, but the end result is always the same — a series of tone-encoded signals that must be detected and decoded to perform the intended function. Over the years a number of articles have appeared in Amateur publications concerning the decoding of the DTMF signal, but what to do with the decoded signals has often remained in question. The circuit described in this article may help to answer that question and perhaps encourage repeater technical committees whose members are thinking about modernizing or expanding their control circuitry.

This multifunctional, expandable decoder features full DTMF decoding, multi-digit coding, latching output, multi-source reset and a separate auxiliary output, for use by system logic. It's DC coupled to greatly reduce falses from noise, and expandable. No adjustments are necessary; no tedious alignment or periodic tweaking are required. Its circuits detect the sequence in which the decoded DTMF signals are received and, if the, sequence is correct, provide a latched open-collector output that can be used to activate the intended function.

circuit description

Two basic circuits comprise the DTMF controller: a DTMF decoder and a sequence detector (see **fig. 1**). For practical repeater control it may be preferable to locate these two circuits on separate PC boards. However, for purposes of explanation, consider them to be located on one subassembly.

The heart of the DTMF decoder is the SSI-201, manufactured by Silicon Systems, Inc. I won't go into details of the chip's operation, except to say that it represents a quantum leap beyond the traditional 567 decoder and exhibits central-office quality with respect to immunity to talk-off and noise falsing. This circuit was described in *ham radio* by KC9C[1] and is essentially lifted from the data sheet* provided by the manufacturer.

One addition that I made was to add the low-pass filter consisting of R1, R2, C1, and C2 (see **fig. 2**), which significantly improved the chip's tolerance to different pads in the field. No problems were encountered when the DTMF tones were generated in a Western Electric Series 35 pad, or equivalent. However, the vagaries of synthesized DTMF generators found in a variety of HTs and DTMF microphones made the decoder a little fussy — some pads would work, while others would not. Since the addition of the LPF, none of the pads used to access the decoder has failed to produce the correct outputs.

When the correct tone pair is received at pin 12 of U1, a hexadecimal output is generated on pins 1, 20, 21, and 22. These four data lines are applied directly

*Data sheet available from Silicon Systems, Inc., 14351 Myford Road, Tustin, California 92680.

By Terry Simonds, WB4FXD, PO Box 1558, Edgartown, Massachusetts 02539

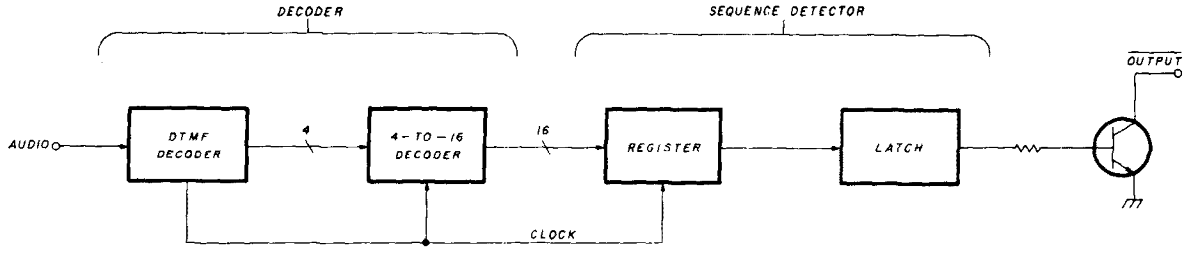
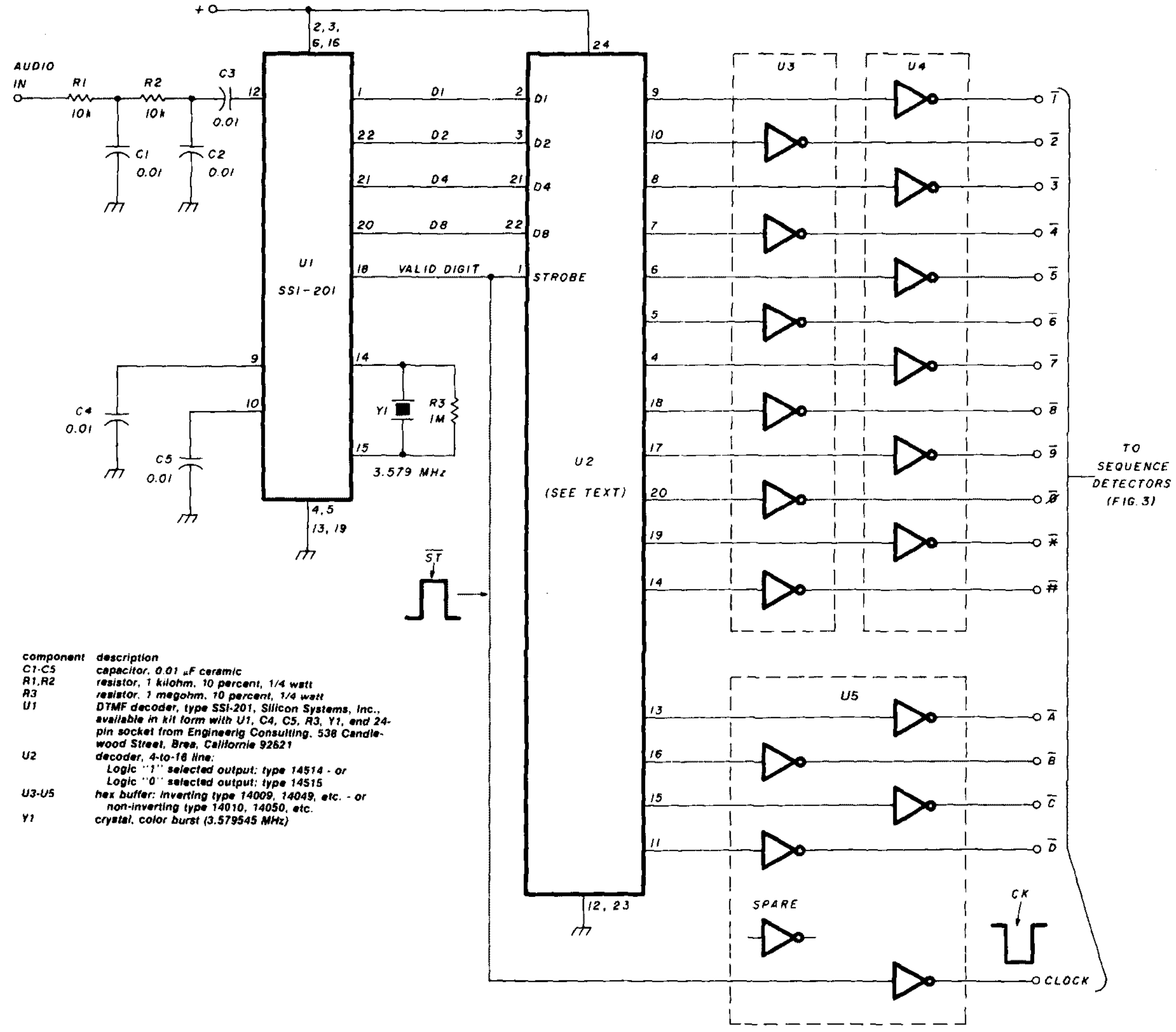


fig. 1. DTMF controller block diagram.



component	description
C1-C5	capacitor, 0.01 μF ceramic
R1,R2	resistor, 1 kilohm, 10 percent, 1/4 watt
R3	resistor, 1 megohm, 10 percent, 1/4 watt
U1	DTMF decoder, type SSI-201, Silicon Systems, Inc., available in kit form with U1, C4, C5, R3, Y1, and 24-pin socket from Engineerig Consulting, 538 Candlewood Street, Brea, California 92621
U2	decoder, 4-to-16 line: Logic "1" selected output: type 14514 - or Logic "0" selected output: type 14515
U3-U5	hex buffer: inverting type 14009, 14049, etc. - or non-inverting type 14010, 14050, etc.
Y1	crystal, color burst (3.579545 MHz)

fig. 2. DTMF decoder schematic diagram. Buffers U3-U5 may be omitted by substituting for U2 (see text).

to U2, a 4- to 16-line decoder. Pin 18 of U1 produces a positive-going "valid-digit" pulse if both DTMF tones have been received and decoded. This pulse is applied directly to U2 as a strobe for that chip's operation and is inverted and applied to the sequence detector as a clock.

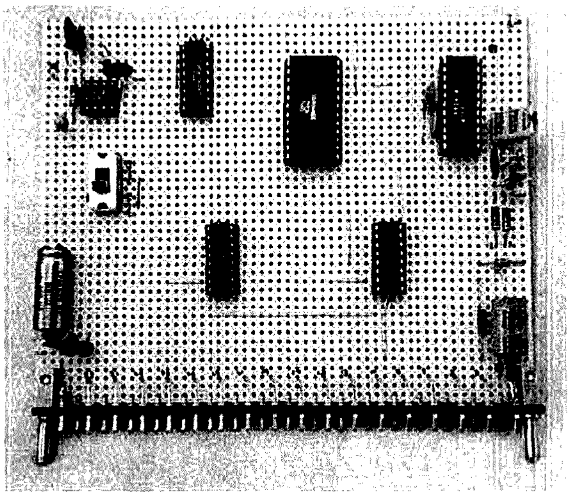


fig. 4. Decoder card. The heart of the decoder is the 24-pin chip — the SSI-201. Several components shown here are not required for basic decoder operation, but are used in other repeater-control functions.

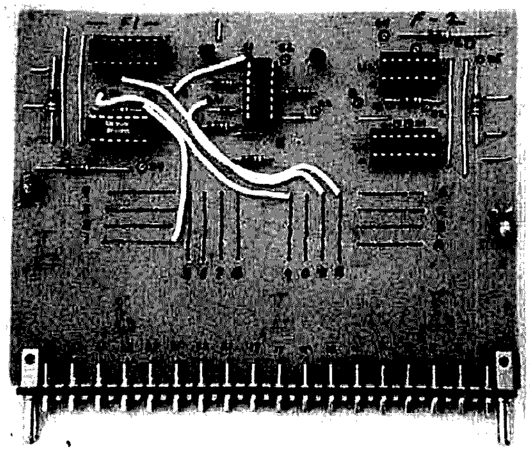


fig. 5. Sequence-detector card. The left half of the card has been programmed with the code *97. The right half is uncoded at the present. The vertically oriented chip at the top center is the output latch shared by the two sequence detectors on the card.

The resulting output from U2 is logic *1* latched on one of its 16 output lines. By appropriately labeling these lines, one arrives at a unique output that corresponds to the button that was pressed on the pad.

Each output line from U2 is buffered and inverted in three hex inverter/buffers, U3-U5, for presentation to the sequence detector that follows. Note that these buffers are shown in the dashed box illustrated in **fig. 2**. When I was designing the control system for our repeater, I was unsure of the number of stages that would follow the decoder, for reasons that will become apparent later. I therefore chose the 14514 decoder, which has a decoded "1" output and inverted this in the buffer. If it's known in advance what the load will be, it may be possible to substitute a 14515 for U2 and eliminate U3-U5 (the 14515 presents a decoded "*0*" on its output line).†

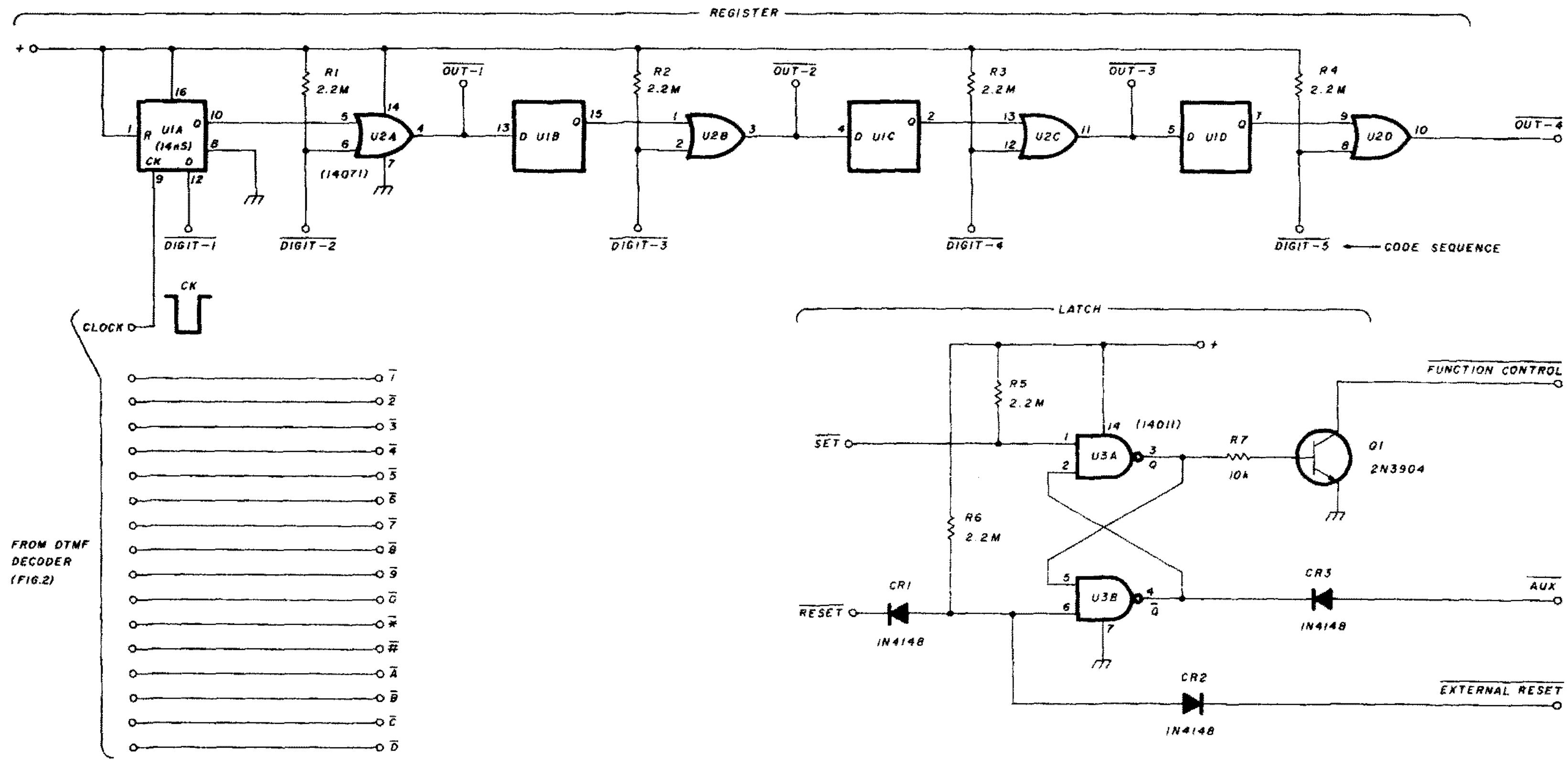
The purpose of the sequence detector is to produce a latched output if, and only if, the sequence of *0*'s received from the decoder coincides with a preprogrammed sequence of digits for the desired function. **Figure 3** shows the schematic diagram of the sequence detector, which consists of a register and a latch. The register, made up of quad-D flip-flop U1 and quad-OR U2, has applied to its $\overline{\text{DIGIT-1}}$ through $\overline{\text{DIGIT-5}}$ inputs the desired decoded digit lines from the decoder in the sequence that corresponds to the desired code. Also applied to the register at U1-9 is the clock signal from the decoder.

As DTMF signals from the repeater receiver are decoded, *0*'s are momentarily placed on various of the 16 lines to the sequence detector. At the same time, a clock signal is generated for each decoded digit. The first digit line in the code is connected to $\overline{\text{DIGIT-1}}$, the second to $\overline{\text{DIGIT-2}}$, and so on. The first *0* received on $\overline{\text{DIGIT-1}}$ is clocked through U1A to its Q output, where it is stored and presented to OR gate U2A. The second decoded digit, presented to the other input to U2A, produces a *0* at the output of U2A which, upon receiving the second clock pulse, is passed on the the Q output of U1B. The third *0* in the sequence is OR'ed in U2B to produce a *0* at U2-3. It can be seen that a logic *0* is then passed along from output to output of the register as the correct code is punched into the pad. Note that there are output lines from the register corresponding to codes of two, three, four, and five digits. Any of these outputs can be used, depending on the number of digits in the code. If an incorrect digit appears in the code, a *1* will be clocked into the register at that point because there would not be a "desired digit line" as an input. This will cause the corresponding $\overline{\text{OUT}}$ line to remain high. The valid sequence is thus lost, and must be entirely re-entered.

The *0* from the appropriate $\overline{\text{OUT}}$ line is applied to $\overline{\text{SET}}$ input of conventional R-S flip-flop U3, configured as a dual-output latch. The Q output (pin 3) saturates transistor Q1 through R7, thereby pulling its collector to ground to actuate the selected function. The $\overline{\text{Q}}$ output from the latch (pin 4) can be used as an auxiliary output if desired. A typical application would be as an interlock signal for the repeater controller logic. If an auxiliary output is not wanted, omit diode CR3.

There are several ways that the latch can be reset and the function turned off; each requires that a logic *0* be placed on the latch $\overline{\text{RESET}}$ line.

†The nominal fan-out capability of a CMOS chip is 50.



component*	description
CR1-CR3	diode, silicon, 1N4148 or equivalent
R1-R6	resistor, 2.2 megohm, 10 percent, 1/4 watt
R7	resistor, 10 kilohm, 10 percent, 1/4 watt
U1	quad-D flip-flop type 14175 or equivalent
U2	quad 2-IN OR gate type 14071 or equivalent
U3(1/2)	quad 2-IN NAND gate type 14011 or equivalent

*per detector — 2 detectors per card.

fig. 3. Sequence detector schematic diagram. Resistors R1-R5 and diode CR3 may not be required; see text.

Single-digit reset. If one of the digit lines from the decoder is connected to $\overline{\text{RESET}}$, the latch will be reset when that digit is received. The most common would be #.

External reset. A logic *0* applied through diode CR2 will reset the latch. For example, if the function selected were autopatch, a negative-going pulse from the autopatch time-out timer could be connected here to reset the latch so that the patch would not be immediately brought back up after time-out. If external reset is not needed, omit diode CR2.

Multi-digit coded reset. If it is desired to provide a coded reset function, such as enabling the repeater transmitter after it has been "put to sleep," then $\overline{\text{SET}}$ is connected to an $\overline{\text{OUT}}$ line higher in number than that used to enable the function. For example, suppose the code *7539 were to be used to disable the transmitter. The five digit lines corresponding to that code would be connected to $\overline{\text{DIGIT-1}}$ through $\overline{\text{DIGIT-5}}$ in the code sequence and $\overline{\text{OUT-4}}$ would be connected to the $\overline{\text{SET}}$ line of the latch. The $\overline{\text{OUT-2}}$ line from the register would be connected to the $\overline{\text{RESET}}$ line of the latch. When the disable code *7539 is received, the latch is set and the output *0* is used to disable the transmitter, for example, by holding the receiver COR line low. To enable the transmitter, the code *75 would be used and the latch would be reset allowing the COR to operate the transmitter. This is not a particularly secure coding system, but it allows turning a function on and off with the same hardware and with a code that the average repeater user would not be able to "guess." Perhaps this is an acceptable tradeoff in component count and reliability.

Another very practical application of single- and multi-digit reset would be in the restriction of certain types of calls made on the repeater autopatch. For toll restriction, use the single-digit approach. Toll calls can only be made by dialing a *0* or a *1* after you have received dial tone. Assume a three-digit autopatch-access code. Connect the digit lines from the decoder in the correct sequence to the sequence detector $\overline{\text{DIGIT}}$ inputs and connect the $\overline{\text{OUT-2}}$ line to the $\overline{\text{SET}}$ line of the latch. Then diode-OR digits *1* and *0* to $\overline{\text{DIGIT-4}}$ and connect $\overline{\text{OUT-3}}$ to $\overline{\text{RESET}}$. When you now try to dial *1* or *0* as the first digit after the patch has been brought up, it will be immediately disconnected. To restrict calls to certain central offices, diode-OR the first two digits of the prohibited offices to $\overline{\text{DIGIT-4}}$ and $\overline{\text{DIGIT-5}}$ and connect $\overline{\text{OUT-4}}$ to $\overline{\text{RESET}}$. This latter method may be particularly attractive to repeater groups whose telephone billing is based on message unit systems.

Note resistors R1-R5. These are pull-ups to hold the gate inputs high if there is not a digit line or output line connected. This is necessary to prevent possible instability due to floating gate input. R6 performs the same function in the latch. Since this gate input is diode-coupled, R6 must be installed at all times.

construction details

The DTMF controller for our repeater was built as several subassemblies; the decoder and its associated circuits occupy one 4.5 × 6 inch (11 × 15 cm) perfboard, while two sequence detectors are located on each of several identically sized PC cards. **Figure 4** shows the decoder card. The 555 timer in the upper left corner is not required for decoder operation. Its function in the repeater controller is to mute the DTMF tones to the transmitter so that they are not retransmitted. This affords increased security of repeater codes, and somewhat calms the nerves of the control operators. The slide switch is used to select muting or non-muting of the tones and is also not required. The transistor, seen near the lower left corner, is not shown on the schematic. It functions as a power-up sequencer for the card, required because of different logic voltages used in the repeater controller logic system, but is not necessary for decoder operation.

Figure 5 shows one of the sequence-detector cards used in the repeater controller. This card is shown with jumpers installed for control code *91 and the diode installed for an auxiliary output for that function. When the photo was taken, the right half of the card was not required and was therefore left blank. Another function has since been added, jumpers have been appropriately arranged, and the necessary chips installed in their sockets.

expansion and change

As every repeater operator knows, there's no such thing as a finished machine. There are always changes to be made, more functions to be added, and codes to change for one reason or another. With the controller described here, this becomes a relatively simple task.

For example, if an access code is to be changed, all that has to be done is to move a few jumpers around. If you want to add a new function and all the sequence-detector cards are in use, it's a simple matter to fabricate another one, code it, and install it in the controller card cage. If the digit and clock lines from the decoder are "daisy-chained" to a reasonably large number of "dedicated-spare" sockets, together with V_{DD}, ground, and reset, then backplane wiring in the card cage at the time of addition would be significantly reduced. All that would be required would be to connect a $\overline{\text{FUNCTION ON}}$ line and the $\overline{\text{AUX OUT}}$ line if desired.

Because, at the time of design, I did not know what

functions were to be included in the final configuration of the repeater, I buffered the digit and clock lines from the decoder card. So configured, the DTMF decoder should be able to drive all the sequence detectors or other loads required by the wildest imagination of any repeater technical committee!

conclusion

The simple circuit described here has been in operation for several months in the Vineyard Amateur Radio Association's 220-MHz machine on Martha's Vineyard. It has performed according to design in all respects and has afforded control of our machine's various functions without fail. Barring a near hit by lightning or a catastrophic failure in the power supply, it should remain operative for years. While not providing the flexibility over control functions that a micro-based system could, this control system comes close and is considerably simpler and less expensive to build.

reference

1. Mark Forbes, KC9C, "A State-of-the-Art Touchtone Decoder," *ham radio*, April, 1983, page 27.

ham radio

 Joe Reisert W1JR

designing and building loop Yagis

Antennas are always an exciting subject. Even more exciting is the introduction of a new type of design.

The loop Yagi is one new design that's become quite a performer, especially on 23 and 13 cm (1296 and 2304 MHz). However, its design parameters have not been widely available (especially in North America) and are often confusing.

Recently I had a call from Vern "Rip" Riportella, WA2LQQ, who wanted to know if the loop Yagi would work well on OSCAR 10, Mode "L." "Of course it will work," I told him, and pointed out that several stations were already using them. "But how do I build one?" he asked. Well, that got things rolling.

At Rip's suggestion, I moved this article a few months ahead on my planned schedule so that more stations could take advantage of the loop Yagi, especially on OSCAR 10. When you finish reading this month's column you should have all the material necessary on loop Yagis performance and know how to design and construct one of your own.

development of the loop Yagi

The loop Yagi, as we know it, came into existence in the back yard of Mike Walters, G3JVL, in 1974. Mike was trying to obtain high gain at 23 cm on a single long boom instead of using a parabolic dish. He tried and discarded many combinations of wire loops similar to quad elements; tuning was always critical and the gain was far less than expected.

Then Mike tried using thin (0.028 inch or 0.7 mm) flat strips of aluminum 0.1875 inch (4.8 mm) wide mounted on a 1/2-inch (12.7 mm) diameter boom. Gain was almost immediately realized. Using the heuristic (cut-and-try!) approach and a large box of different length loops, he moved the elements around to various positions on a boom to optimize gain as indicated on a signal strength meter.

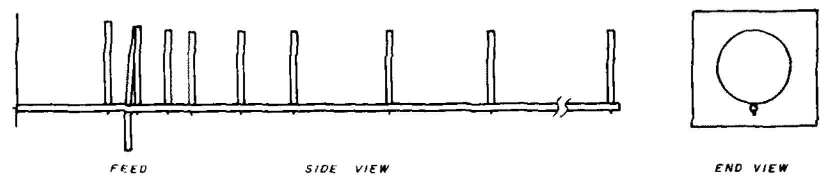


fig. 1. The original G3JVL loop Yagi. Note the plane reflector. The feed passes through the boom rather than through the driven element.

What finally evolved was a 24-director loop Yagi on an 81-inch (206-cm) boom with one driven element and two reflectors, as shown in **fig. 1.**[1] One reflector was a regular loop placed close behind the driven element. The second reflector consisted of a rectangular perforated aluminum sheet measuring 4-1/2 by 5-1/2 inches (11.5 × 14 cm) and mounted on the rear of the boom. The driven element is fed directly at the top with a short length of 0.141 inch (3 mm) diameter 50-ohm semi-rigid coax.

Yagi versus loop Yagi

You may ask, "Why all the fuss about loop Yagis?" The answer lies mainly in their mechanical design. Really long Yagis (over 10 wavelengths long) with high gain (over 18 dBi) were not readily available until the 1980's.[2] Furthermore, long Yagi dimensions are *very* critical if optimum gain and low side lobes are desired. UHF and microwave designs are even more critical. For instance, at 23 cm the typical Yagi tolerance should be within ±0.003 wavelengths, with 0.001 wavelength preferred.[3] This equates to a 23 cm tolerance of 0.027 and 0.009 inches (0.069 and 0.023 mm), respectively — and these tolerances are quite difficult to maintain.

Metallic boom Yagis also require a correction factor to be applied to each element. (This area is still quite controversial.) Add this to the problems of conductivity to the boom or the capacitive and inductive effects of through the boom mounting with insulated elements and you have a tall order to fill.

Don't overlook the feed system in a Yagi *antenna*. For a good match, you must have a good mechanical assembly that can tune out variations in the design, and the antenna should preferably be fed with a balanced feed system. This implies the use of a *balun*, another item requiring good tolerances and construction techniques.

Many of the previously mentioned items were designed out or improved in the G3JVL loop Yagi designs. Mechanical duplication is straightforward. For instance, elements are mounted on top of the boom and connected with just one screw. Various boom diameter corrections are quite well documented, as we shall shortly see. Similarly, the loop length, width and thickness have been carefully tabulated with correction factors for different sizes and frequencies. Finally, a unique driven element feed system with a built-in balun was developed. Thus the loop Yagi is relatively easy to duplicate.

By now you must be asking what the disadvantages must be. Well, the main problem is that there just aren't a lot of recipes. In fact, there have

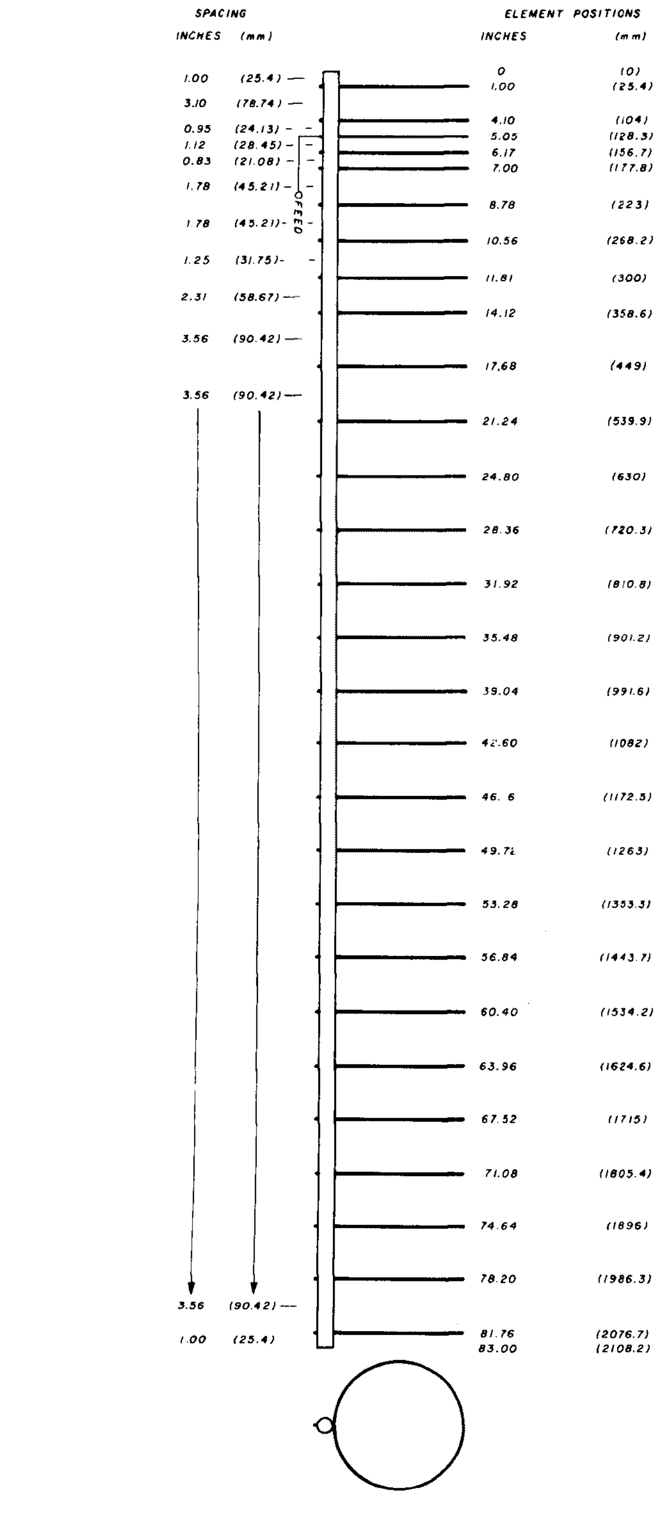


fig. 2A. The basic 28-element loop Yagi, with mechanical dimensions based on 1296 MHz. When constructing a 38 or 45-element loop Yagi, add additional loops as required, spacing them each 3.56 inches (90.42 mm) further down the boom as described in text.

been only three major designs published for 28, 38, and 45 elements.[1,4,5] Even these designs could probably use some improvements to reduce the first side lobe levels.

The loop Yagi also has a fair wind load. Several years ago Rick Commo, K1LOG, and I set out to build a 70 cm (432 MHz), 28-element loop Yagi. It looked truly inspiring and had very good gain when we measured it on my backyard antenna range. But plans to install it at K1LOG's QTH had to be abandoned halfway up the tower, because even in a 10 mph (16 km) wind, holding on to the antenna was like holding on to a large sail! So loop Yagis simply aren't practical below about 900 MHz.

Another loop Yagi headache is large birds since they can easily bend or flatten the loops if they decide to roost on them. G3JVL solved most of this problem by mounting his loop Yagi antennas upside down!

A third problem is deterioration of the joints after long exposure to the elements. This, however, is quite manageable, as we shall see.

loop Yagi variations and improvements

Many of the disadvantages mentioned above have been solved or circumvented and the loop Yagi — still a high-performance antenna — is widely accepted on OSCAR 10 and the 23 and 13 cm bands. If you're looking for an antenna that's relatively easy to duplicate and provides high gain on 23 cm and up, this may be the one for you.

Soon after the original design was published, several persons built models and frequency-swept the response.[1] Some dips in gain were noted, especially below the frequency of interest. Although this isn't normally a problem, it can be if tolerances vary. The first improvement was the addition of another director between directors 4 and 5 of the original design.[4] This also broadened bandwidth and increased gain by about 0.5 dB.

Shortly after I built my first loop Yagi

in 1978, I became disillusioned with the plane reflector. It was difficult to fabricate and has a wind load that caused it to act like the tail end of a weather vane! But removing the plane reflector caused the gain to drop as much as 1 dB. I tried a variety of substitutes and finally settled for a second reflector loop, identical to the first one (just behind the driven element) but placed at the rear of the boom. It had far less wind load than the plane reflector and was easier to fabricate; the only perceptible change noted was perhaps a 0.1 dB drop in forward gain, a small penalty for the simplification it yielded.

Since the original design was conceived, the feed system has been improved. Originally the semirigid coax passed to the rear of the driven element and through the boom as shown in **fig. 1**. G3JVL *found that if you* pass the same coax directly through the lower portion of the driven element and correctly shape the element, a quarter-wave type of balun is formed.[4]

Other improvements made by G3JVL included higher gain for the original design, longer boom models, and correction factors for changing loop thickness and width as well as boom diameter. The latter makes frequency scaling the loop Yagi a breeze.[5]

how to design a loop Yagi

Let's look at the designs available. The improved original design now uses 28 elements with two different director lengths as shown in **fig. 2** and **table 1**. *All the loop lengths shown in* **table 1** *are based on the original design with a boom diameter of 1/2 inch (12.7 mm) and loops made from flat aluminum strips 3/16 inches (4.76 mm) wide and 0.028 inches (0.71 mm) thick. This will be the reference point for all other designs.* Also note that the length of the loops is equal to the distance between the mounting holes. An extra 0.25 inch (6 mm) should be added to each end.

I immediately found these sizes to be unacceptable. First, the elements were too narrow and would often break at the screw holes when they were tightened to the boom. This problem was overcome by using a wider (0.25 inch or 6.35 mm) strip. Also the 0.028 inch (0.71 mm) aluminum is not commonly available in the United States but 0.032 inch (0.81 mm) material is. Finally, the 1/2 inch (12.7 mm) diameter boom is too flimsy in areas where snow and ice are common. Adding extra support struts helps, but a larger diameter boom is still recommended.

Calculating these changes becomes very simple with the correction factors that G3JVL later derived.[5] This information is shown on individual graphs in **fig. 3** (the correction factor for boom diameter changes) and **figs. 4** and **5** (correction factors for different loop widths and thicknesses, respectively).

Let's work through an example. I prefer to use a 3/4 inch (19 mm) boom and loops that are 1/4 inch (6.35 mm) wide and 0.032 inch (0.081 mm) thick. Looking at **figs. 3**, **4**, and **5**, we see that corrections are +0.9, −0.3, and +0.04 percent, respectively. Adding these corrections together, we get a net change of +0.64 percent. This means that if we select new sizes of materials, all we have to do is increase the sizes of all directors and reflectors shown in **fig. 2** by 0.64 percent. Therefore, the scaled directors will now be 8.303 inches (211 mm) and 8.051 inches (205 mm), respectively. The reflectors will measure 9.732 inches (247 mm).

other designs

The compact 28-element has a gain equivalent to that of a 35-inch (90-cm) dish. But if you want a model with a higher gain, you may want to build a 38 or 45-element loop Yagi — roughly equivalent to a 39-inch (1 meter) and 43-inch (1.1-meter) dish, respectively — instead.

Table 1 shows the lengths of the 38 and 45-element designs. *Again, these are based on the original boom diameter and loops as noted above.* Note

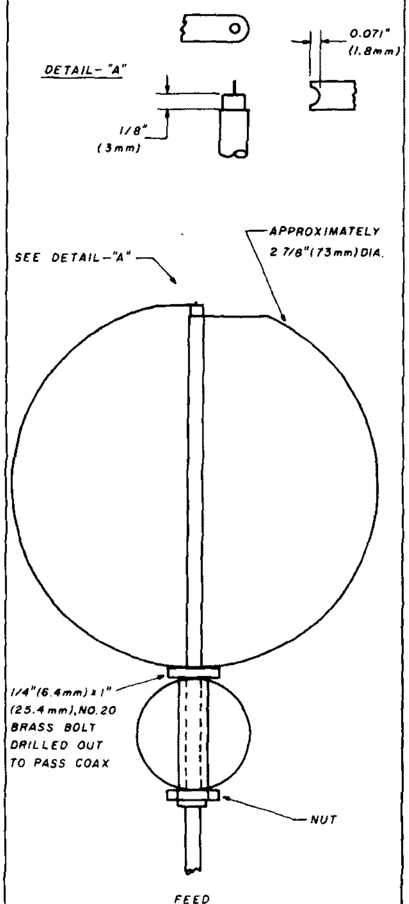


fig. 2C. The driven element is constructed as shown using a brass or copper strap 0.25 inch (6.35 mm) wide and 0.032 inch (0.81 mm) thick. The length is shown in *table 1* and is the distance between the center of the attachment holes. See text for final checkout.

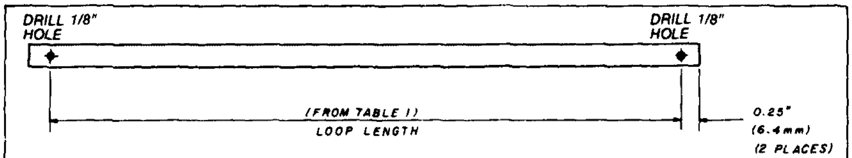


fig. 2B. Loop length is measured from hole to hole. Note that an extra 0.25 inch (6.4 mm) of aluminum strap should be provided at both ends of each loop for mechanical stability.

table 1. 1296 MHz loop-Yagi element lengths.

design	28 elements inches	28 elements mm	38 elements inches	38 elements mm	45 elements inches	45 elements mm
reflector 1 & 2	9.67 (9.732)	246 (247)	9.67 (9.732)	246 (247)	9.67 (9.732)	246 (247)
driven element	9.23 (9.292)	234 (236)	9.23 (9.292)	234 (236)	9.23 (9.292)	234 (236)
directors 1-11	8.25 (8.303)	209 (211)	8.25 (8.303)	209 (211)	8.25 (8.303)	209 (211)
directors 12-18	8.00 (8.051)	203 (205)	8.00 (8.051)	203 (205)	8.00 (8.051)	203 (205)
directors 19-23	8.00 (8.051)	203 (205)	7.70 (7.749)	196 (197)	7.75 (7.800)	197 (198)
directors 24-25	8.00 (8.051)	203 (205)	7.70 (7.749)	196 (197)	7.65 (7.699)	194 (196)
directors 26-35			7.70 (7.749)	196 (197)	7.65 (7.699)	194 (196)
directors 36-42					7.50 (7.548)	191 (192)

The numbers shown above are the reference element lengths for designing a loop Yagi at 1296 MHz. These dimensions are based on a 0.5 inch (12.7 mm) boom diameter with 3/16 inch (4.76 mm) wide by 0.028 inch (0.71 mm) thick loops. The lengths shown in parenthesis are for use of a 3/4-inch (19 mm) boom and loops that are 1/4 inch (6.35 mm) wide and 0.032 inch (0.81 mm) thick. Other size material can be used if scaled as discussed in the text.

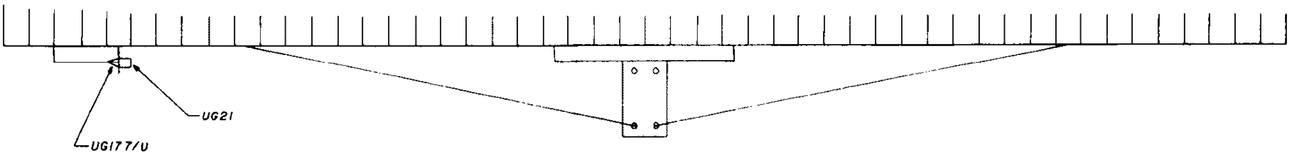


fig. 2D. Overall assembly showing connector mounting, mast bracket (per *fig. 7*) and the recommended supporting bars. The feed system shown may not be desired if stacking is used as described in text.

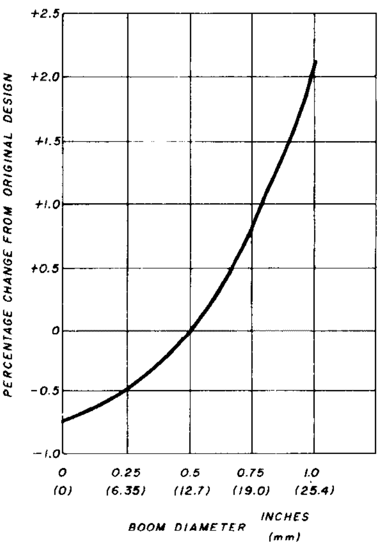


fig. 3. Use this graph to determine correction factors for booms other than 0.5 inch (12.7 mm) in diameter.

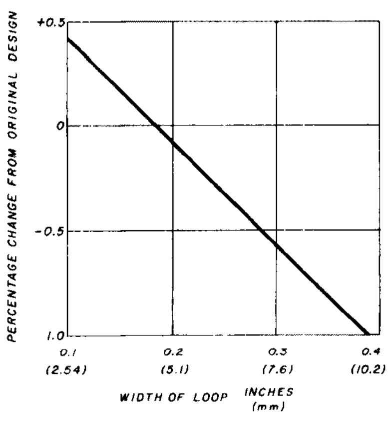


fig. 4. Use this graph to determine correction factors for loop widths other than 0.188 inch (7.76 mm).

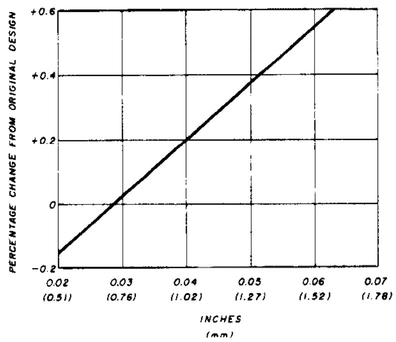


fig. 5. Use this graph to determine correction factors for loop thickness other than 0.028 inch (0.7 mm).

that all new directors on the longer boom designs are moved out an additional 3.56 inches (9.04 cm). However, the diameter of the loops varies with a slightly different tapering schedule.

For those who don't want to go through the scaling exercise, I've listed, in table 1 (in parentheses), the recommended sizes for the 28, 38, and 45-element designs using a 3/4 inch (19 mm) boom and loops that are 0.032 inch (0.91 mm) thick and 1/4 inch (6.35 mm) wide. Different sizes can also be used, but you'll have to work out the new values yourself!

frequency scaling

If you want to operate at a frequency other than 1296, how do you perform the scaling? It actually isn't that complicated. All it requires is a hand calculator, some scratch paper and some thought! All data can be found on the scaling charts.

For example, suppose you want to make an optimized OSCAR Mode L design. 1269 MHz is actually only

table 2. Specifications and stacking information for the 28, 38, and 45-element loop Yagi designs discussed in the text.

design	boom length (in λ)	gain* (dBi)	equivalent dish size* (inches)	(meters)	BW* (degrees)	stacking distance in λ (E plane)	(H plane)
28-element	9.0	19.0	35	0.9	18.0	2.50	2.30
38-element	13.2	20.0	39	1.0	16.5	2.75	2.55
45-element	15.7	20.7	43	1.1	15.0	2.85	2.65

*Approximate

about 2 percent lower in frequency that 1296 MHz. In reality, the models shown work quite well even though they're made for 1296 MHz, since the bandwidth of a typical Yagi type structure is usually good for at least −2 to +1 percent. But let's say you really want to optimize the design without compromise.

In order to exactly duplicate the performance, *all element dimensions and spacings must be increased by 2 percent* (1296-1269/1296). The spacing is easy. Just make all spacings 2 percent or 1.02 times the values shown in the figures and the table.

Then scale the new boom diameter. Let's say you want to use a 3/4 inch (19 mm) boom at 1269 MHz. First you have to design the boom as if it were 2 percent *smaller in diameter at* 1296. For example, if we have a boom 0.735 inches (18.67 mm) in diameter at 1296, it will be equivalent to 0.750 inch (19 mm) in diameter at 1269 MHz. The correction factor for 0.735 inch (19 mm) at 1296 is approximately +0.8 percent.

The elements are done in the same fashion; they're shortened by 2 percent at 1296 MHz and looked up in **figs. 4** and **5**. Then the changes are added to the boom correction for the final correction. If the frequency is above 1296, the boom correction and element parameters must be lengthened by the equivalent percentage at 1296. Proceed in the same way as already described.

construction

The loop locations should be marked carefully on the boom as shown in **fig. 2**. *This is a very important procedure that should not be taken lightly*. Because spacings are quite close for the first directors, a build-up tolerance could occur. I've therefore shown all dimensions referenced from the rear of the boom to prevent tolerance buildup.

For best symmetry, the boom should be drilled while supported in either a "V" block or a Porta Vise.™ After the first element hole is drilled, a dummy element should be fastened in place. This will provide a reference point for drilling successive holes so that they'll all be located in the same plane. Admittedly, this is less of a problem on a loop Yagi than a conventional Yagi since the elements can be leveled by pushing them sideways for alignment. Finally, drill out the larger hole for the driven element mounting screw.

Next, center-punch each loop for the proper length as shown in fig. 2, leaving about 0.25 inches (6.35 mm) of extra material on each end. Make sure the punch is centered on the loop width since any sideways skewing will weaken the joint. Also, use the smallest possible drill to just pass the screw shank.

The loops can be pre-formed by rolling them on a piece of tubing or dowel. Next insert the screw while drawing the two ends together, inserting the screw through the boom in the correct location and securing it with a lock washer and nut. *Do not overtighten; this will deform the loop material.*

Assembling the driven element is a little tricky. It must first be soldered together as a sub-assembly as shown in **fig. 2**. Then pass the mounting screw through the boom and secure it with the nut. (Suitable semi-rigid transmission line can be purchased from Microwave Components, WB8EUU, 11216 Cape Cod, Taylor, Michigan 48180, for approximately $2.00 per foot.)

Now add the connector. For maximum strength, rigidity, and reliability, it should be mounted on a small right angle plate attached to the boom. This simple, low-cost assembly with the shield, suggested to me by Bob Johnson, K9KFR, has worked well on the four loop Yagis I presently use.

Alternatively, a type OSN™* male "N" cable connector can be attached directly to the semi-rigid transmission line. This method offers an advantage if you're going to stack your antennas. You can use a transmission line the proper length required to reach the power splitter (more on this later).

other variations

Other loop widths can be used on this design if they're properly scaled by the methods already discussed. For instance, if some of the elements are widened, it may be possible to use a uniform length loop on the shorter model.

As mentioned earlier, the simplicity of the element attachment to the boom is one of the selling points of this design. But the place at which the loop attaches to the boom is a high current point. Therefore, if the elements are not tightened properly, the joints can corrode, resulting in increased loss in the future.

Several methods have been used to circumvent this problem. One is to coat the joints with an epoxy or sealant type of material after assembly. A second, more complex mounting method is to rotate all the elements 90 degrees.

First, a single screw, washer, and

*"OSN" is a trademark of Omni-Spectra, Inc.

nut hold the loop ends together. Then a second hole is drilled one-fourth the distance between the end mounting holes in each loop. Finally the loops are attached with another screw to the boom. This locates the point at which the loops are attached at a voltage node.

Remember that all sizes given, including sizes on the scaling charts, are based on mounting loops exactly as shown. Any deviations will upset the correction factors and possibly degrade performance. Several years ago, for example, a number of these loop Yagis were distributed in kit form by a midwest Amateur. Although the design was satisfactory, the instructions recommended, for mechanical reasons, that a lock washer be placed between the loop and the boom rather than between the boom and the nut. The effect of this instruction was subtle. Since the element was now further removed from the boom, the boom correction was diminished, and the elements therefore resonated at a slightly lower frequency. The net result was a 0.5 to 0.75 dB gain decrease and higher side lobes. Removing the washer, cleaning up the surfaces, and then placing the washer on the other side of the boom instantly restored the antenna's original design performance.

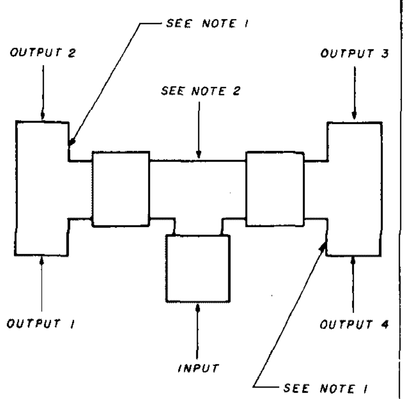


fig. 6. This is a simple power splitter using three type "N" coaxial Tee adapters. *Note 1:* Type UG 107A/U recommended but UG 107B/U is usable with slightly higher VSWR. *Note 2:* Either a UG 107B/U or UG 28A/U are usable.

stacking loop Yagis

The designs given in this month's column will work fine by themselves. However, there's always somebody who wants more gain. Since loop Yagis have lower wind loss than dishes (one of the prime reasons for using this design), they're often stacked to obtain higher gain.

Recommended stacking distances and configurations are discussed in references 6 and 7. **Table 2** lists electrical parameters to follow for proper stacking on all three models just discussed. Generally speaking, the spacings shown may seem too close, but they were confirmed at my QTH by a series of tests using both a signal source and on-the-air test.

When stacking antennas at such high frequencies, phasing harness losses and phase lengths are very important. Four matched phasing lines can be brought to the mast where a power splitter is attached — or you can use an identical length of the same amount of semi-rigid coax as mentioned earlier. You can then use short phasing lines with the back plane feed method as discussed.

Any suitable two-way or four-way in-phase power divider can be used.[7] One simple and usually inexpensive power divider can be made for 23 cm operation by combining three UG 107A/U type N Tee adapters as shown in **fig. 6.**[8] The longer (and older) type UG 107A/U is recommended because it's closer to the desired phase length. However, the newer UG 107B/U will work with slightly higher VSWR. These coax adapters are readily available at flea markets.

Loop Yagis are usually mounted on brackets similar to the one shown in **fig. 7.** These brackets are then attached to a mast or a stacking frame. If two- or four-loop Yagis are stacked vertically, it's often more convenient to turn the lower ones upside down so that the mast doesn't have to pass through the boom or loops. In this case, remember to invert the feed on the upside down antennas so that they're not fed out of phase.[7]

You may ask whether the mast can be placed through a loop Yagi as is typically done on Yagi antennas. The answer is yes, but the diameter should be kept reasonable and the mast should preferably pass midway between two loops. Tests conducted by G3JVL showed a slight gain reduction

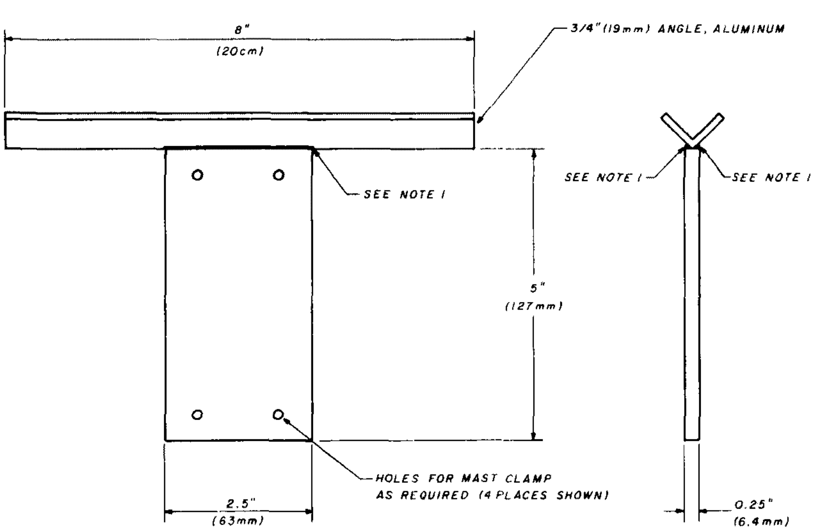


fig. 7. A recommended mast to boom mounting bracket. Material is aluminum. *Note 1:* Heliarc the angle aluminum and plate together at this point. Most local welding shops can provide this service.

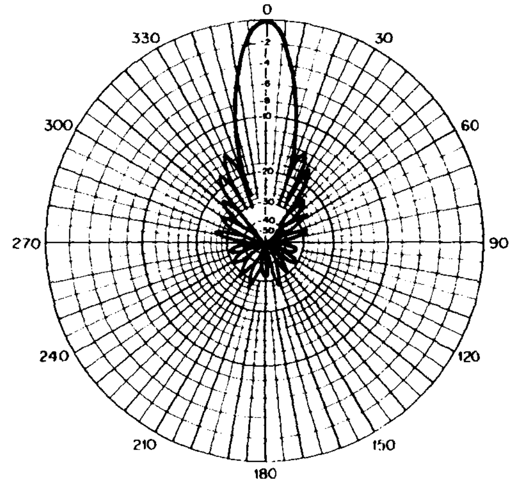


fig. 8. This is a polar plot of the typical "E" plane of a 45 element loop Yagi. Other designs will have similar patterns but with wider half-power beamwidth as shown in *table 2*.

(approximately 0.25 dB) when a large (2 inch or 51 mm) diameter boom was passed through the plane of a 23-cm loop Yagi.

Supports or trusses should be used to stabilize and strengthen the boom, especially where ice and snow are prevalent. This can easily be done by attaching a piece of tubing between a point about two thirds the distance out on the boom and the bottom corner of the mounting plate shown in **fig. 7**. This tubing can have the same or slightly smaller diameter as the boom.

final performance evaluation

Once the loop Yagi is completely assembled, it should be mounted on a mast in a clear area and pointed away from any local objects. Then the VSWR should be measured. If it is not below 1.5:1, bend the first reflector or the first director closer or further away from the driven element for a better match. Sometimes even the second director can be bent slightly to enhance VSWR., If all else fails, shorten or lengthen the driven element slightly.

After the antenna is mounted in its final resting place, the pattern can be tested. The beamwidth is narrow as shown in **table 2**. The first side lobes should be at least 10 dB and usually 13 or so dB below the main lobe. If not, you may have scaled your dimensions improperly.

For reference, see **fig. 8**, in which a pattern adapted from an equivalent 13 cm (2304 MHz) 45 element loop Yagi is shown. The initial data for this pattern was done on a professional antenna range and supplied to me by N3CX. The 2304 MHz loop Yagi will be covered in a paper I'll be presenting at the First Annual 1296/2304 MHz conference in Estes Park, Colorado from September 20 to 22 (see end of column). Note that the nulls between the main beam and the side lobes can be easily used to determine beamwidth, as described in reference 9.

summary

The loop Yagi is an exciting and relatively new antenna design that has been well received as a true performer. In this month's column I've tried to provide some overall background information on its design along with procedures for modifying the design for other frequencies and material sizes to suit the builder.

This antenna isn't a panacea, nor does it necessarily offer any more gain than a well-designed Yagi. However, it's easily reproduced and has much less wind load than an equivalent dish *of the same gain*. Models have been made to frequencies as high as 3 cm (10 GHz) with equivalent performance.

acknowledgements

This design would not have been possible without the many hours of trials and testing by Mike Walters, G3JVL. I'd like to thank him for all the encouragement and information supplied to me over the years so that I too could obtain the gain claimed. Also, thanks to N3CX for providing the pattern data of the 45-element loop Yagi shown in **fig. 8**.

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important VHF/UHF events

September 7-8:	*International Region 1 VHF Contest*
September 14-16:	ARRL VHF QSO Party
September 16:	EME Perigee
September 20-22:	*First Annual 1296/2304 MHz Conference, Estes Park, Colorado (Contact WØPW)*
October 5-6:	*Mid-Atlantic States VHF Conference, Warminster, Pennsylvania (Contact WA2OMY)*
October 5-6:	*International Region 1 UHF/SHF Contest*
October 9:	*Peak of Draconids Meteor Shower predicted at 0300 UTC*
October 15:	*EME Perigee*
October 20:	*Peak of Orionids Meteor Shower predicted at 1100 UTC*

ham radio

PRACTICALLY SPEAKING . . .

by Joe Carr, K4IPV

now that the warranty has expired. . .

The big day has finally arrived. After months — or years — of scrimping and saving, you've plunked down a kilobuck or so for a new transceiver. It's a gleaming engineering marvel with knobs, dials, and a luminous digital display. A firm squeeze on the Push-To-Talk button and your voice projects instantly and effortlessly around the world.

But sooner or later, long after the initial euphoria has worn off, you're going to have to consider what we all dread: repairing the rig. Repairing electronic equipment is like buying insurance; most of us never give it a moment's thought until something catastrophic happens. A little planning in this regard can make the difference between quick success or being off the air for weeks while the rig moves through the repair "pipeline."

The first time to think about repairs is while you're unpacking the new rig: should it have to be shipped back to the manufacturer's or importer's repair shop, that original packing carton is probably the best shipping container available. Unpack carefully, taking care not to destroy the carton in your eagerness to get to the goodies.

You think that's a bit extreme? After all, who has room to store that enormous cardboard shipping crate? I called the repair shop service manager for a major mail order seller of ham gear; he told me that all too often, improper packing results in shipping damage, often hidden, that must be repaired at the owner's expense *before* the warranty problem can be examined. Sometimes the shipping damage completely masks rightful warranty repairs — for example, a broken printed circuit board track.

An old grocery carton and three inches of newspaper are not proper shipping materials! Nor is it prudent to use a single wrap of masking tape to seal the carton; rigs have been "lost in the mail" because of inadequate sealing of perfectly adequate shipping containers. Use *two* runs of broad nylon filament tape, wrapped around the carton in both directions. Top this off with plastic film tape or paper packing tape (the kind you have to moisten). Tape a card with your name, callsign, address, and telephone number directly to the rig, just in case your rig becomes separated from its packaging.

It's not prudent to ship more than one item per container. My contact told me of one fellow who shipped an in-warranty Kenwood TS-120S in the same substandard carton as the 12-volt DC, 20-ampere AC power supply that powers the rig. That heavy power supply bounced around that newspaper-lined cardboard box like a cannonball and mashed the TS-120S to bits! Unfortunately, because the damage was caused by improper packing, neither the shipper nor the manufacturer could cover the damage.

In general, it's wise to not ship items such as external power supplies, microphones, telegraph keys, loudspeakers, headphones and other accessories unless told to do so by the repair shop. Twenty years of working in various repair shops left me with many memories of lost customer property. If your transceiver is bad, then ship *only* the transceiver. Keep the other stuff at home. You can be certain that an authorized repair shop will have an adequate DC power supply, speaker or microphone on hand for testing the equipment.

do you really have a problem?

In one shop where I worked, our service jobs ran about 40 percent "NFF" — No Fault Found — although some of the technicians had a more vulgar way of putting it. That means that the user diagnosed a fault when none existed, and the unit was out of service for no reason at all.

Eliminating the NFF takes a little common sense. If you suspect a malfunction, check all connections, accessories, and switch settings. Consult the troubleshooting chart in the owner's manual. I know it sounds dumb, but many a "probable blown fuse" complaint results from the AC power cord's being disconnected from the wall socket.

You'd be surprised to know how often a "won't transmit" complaint is traced to a broken PTT wire in the microphone connector. If your unit won't transmit, try several operating modes. If the rig won't "push to talk," but does work on CW mode, then it's a fair bet that the microphone is faulty. Similarly, if your rig will work on VOX but not on PTT, then the PTT wire is probably at fault. A little common sense goes a long way.

I prefer to keep certain accessories on hand to check my rig; for example, I normally have a spare microphone. Because I'm rarely fond of the standard handheld microphones that come

with new rigs, I buy another model — either a desk mike for the home rig or a touchtone mike for the 2-meter mobile rig. It's a simple matter to squirrel away the original mike, or I would buy a standby replacement mike, to use in troubleshooting problems later.

Other useful items to have on hand are a dummy load and RF wattmeter. Although I now use a Bird Model 43, a simple $30 VSWR meter is also handy for troubleshooting. In the latter case, however, the reading is relative only, so it helps to make a record of readings and knob settings when the rig is known to be working properly. Take the readings with the rig driving a dummy load, and then record the readings and the knob settings that created them. A deflection of 50 to 80 percent of full scale is suitable.

what to fix

Although some "old timers" like me, with 25 or more years in Amateur Radio behind us, loathe the idea of not being able to repair our own rigs, it's nonetheless necessary to recognize that some problems will inevitably be "Beyond Capability of Maintenance" (BCM). If the rig is BCM, then it needs to go the service shop.

You have to decide what repairs you're willing (and able) to undertake. Part of this decision will depend on the design of the rig, while part will be based on the capabilities of your workshop test equipment. Another factor is your personality: if you're fearful of breaking into your rig, then it's probably best to leave that job to others.

Of course, it's also possible to reduce the fear factor by working under the supervision of an "Elmer" who has repair savvy.

In general, the repair of purely mechanical problems with minor controls (for example, toggle switches or potentiometers), visible solder joint problems on printed circuit boards, and DC power supply problems should be well within the capability of most Amateurs. It's also possible to add digital displays to the list above. If the design of your transceiver is such that most assemblies are on removable PCBs, then swapping boards is also on the list of "can do's."

Problems with electronic equipment tend to recur in individual sets within a model group. Because professionals see the same problems again and again, they soon develop a "sense" for the symptoms. During my college years, I worked part-time in an auto radio shop that supposedly didn't hire part-time help. I'd gone into the shop to talk to the owner, who tried to ignore me while he worked on a Delco radio. As we talked, I recognized the problem and told him — without touching the radio myself — to go get a 7281933 trimmer capacitor and replace the LO capacitor. My point in telling this story is to encourage you to call the repair technicians at the manufacturer's or importer's facility, describe the symptoms, and take their advice. Whatever your problem is, they've probably seen it before, and they'll most likely be willing to help you.

While you have a technician on the line, ask whether there are any updates, retrofits, or engineering changes that would normally be incorporated into your rig if it were on the bench right now. You'd probably be surprised to know how often manufacturers and importers will update equipment returned for repair at little or no cost to the owner. Such a courtesy can often eliminate future problems.

tools and supplies

If you don't have a certain minimum assortment of tools, don't even think about attempting the repair of your radio. If you decide to proceed, perhaps the first thing you'll need will be the screwdrivers required for removing the covers. Be sure to have the correct size screwdrivers; improperly sized screwdrivers will tear up both the screw heads and the screwdriver — or pop out of the heads and scratch the cabinet's paint job.

Many rigs are made in Japan. Because the bevel angle of Phillips screws on many Japanese products differs from that of American-made Phillips screws, American-made tools will turn improperly inside the screw slots and tear up the screws. This is one instance in which cheap two-for-a-buck imported "tool barrel" screwdrivers work better than classy $3 models.

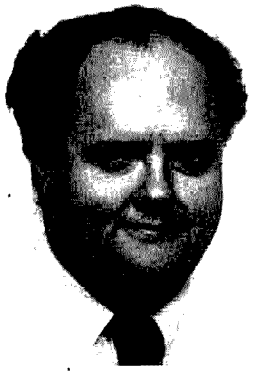
You'll also want to have a couple of soldering devices on hand. For small work (and PCB repairs), use a 25-watt to 75-watt pencil iron; for larger jobs (and wire antennas), use a 250-watt soldering gun. (My own *Weller D-440* has served for 20 years.)

The choice of solder and desoldering aids are as important as the soldering iron you select. The only type of solder to use is *resin core* solder marked for radio, TV, or electronic use. I know many experienced hands will smirk, claiming that this advice is a useless restatement of the obvious, but it's necessary nonetheless: every shop I know occasionally sees a set that is ruined from the use of the wrong solder. Don't use an acid core solder, "plumber's solder" or "industrial solder." All these products use an acid core that will corrode your equipment into the junkyard! Use only resin core solder marked for radio, TV, or electronic uses. (I prefer either *Kester* or *Ersin Multicore* brands in either 50/50 or 60/40 lead/tin mixtures.)

Solder size is also somewhat important. For light work on things like IC pins or other PCB points, I prefer a No. 22, No. 24, or even No. 26 solder. Larger work or wire antennas require a No. 14 to No. 20 size. I keep rolls of No. 14, No. 18, and No. 24 solder in my toolbox. Solder isn't cheap, by the way — so either learn to use it sparingly or get rich and foolish.

Desoldering is a workbench skill that is just as important as soldering, but is all too often ignored in Amateur circles. Several aids available for desoldering, especially on printed wiring boards: desoldering tips, solder suckers, and solder wick.

• **Desoldering tips** are special soldering iron (or gun) tips that are shaped like the pin pattern for the device being



meet Joe Carr, K4IPV

Licensed since 1959, Joe Carr has written more than 221 magazine articles and 41 books on a wide variety of subjects including Amateur Radio, computers and computer programming, electronics, mathematics, and biomedical equipment repair.

Joe began his lengthy and diversified career as a techician, working in mobile and biomedical electronics, communications, and both radio and television broadcasting. He won ISCET certification in consumer and communications electronics in 1971 and 1973, respectively, and was co-recipient, in 1977, of that organization's Technician of the Year Award.

He holds an MSEE from George Washington University (1981) and is currently employed as an electronics engineer in avionics.

Joe's recent books include *Interfacing Your Microcomputer to Virtually Anything, Designing and Building Electronic Gadgets (with Projects), 104 Weekend Electronics Projects, CMOS/TTL: A User's Guide (with Projects),* and *The Complete Handbook of Radio Transmitters,* as well as *The Complete Handbook of Radio Receivers,* all published by Tab Books, Inc. Other titles — published by Reston Publishing Company, a division of Prentice-Hall — include *Elements of Microcomputer Interfacing, Designing Microprocessor-Based Instrumentation, and Elements of Electronic Instrumentation and Measurement.*

Five additional books have been completed and are awaiting publication, and six more are "in the works."

ham radio welcomes K4IPV aboard.

desoldered. For example, there are tips that fit over all 14 or 16 pins of a DIP integrated circuit. The tip is chucked up in a soldering gun so that all pins can be heated simultaneously.

- **Solder suckers** are vacuum tools that, when used in conjunction with a soldering iron, suck melted solder from a joint. Really fancy (high dollar) systems are available; these use a vacuum pump and a pneumatic plumbing system that's an integral part of the soldering iron. Such tools are suitable for commercial shops or industrial users. Amateurs and most other users can get away with less elaborate solder suckers.

There are at least two types of solder suckers that are practical for Amateur applications. One type is a rubber squeeze ball (something like those little ear irrigating syringes you can buy at the drug store) with a nylon or teflon desoldering nozzle. The other type of solder sucker is a spring-powered piston device. The operator cocks it by engaging the spring, then places the tip against the molten joint and presses the spring release trigger. The solder zips up into the tool, leaving a clean joint. If you buy one of these devices, be sure to also buy a spare tip. Worn out tips tend to splash solder around the PCB a little bit. A small file or penknife does wonders in repairing worn-out tips, but a spare tip is nice to have on hand.

A small collection of other hand tools is also necessary. You should have a small pair of diagonal sidecutters, small longnose pliers, mid-size longnose pliers, wire strippers, several sizes of nutdrivers, and some alignment tools suitable for the type of radio set that you own. With this last comment I feel compelled to issue a stern warning: ALIGNMENT ADJUSTMENTS ARE *NEVER* USED IN TROUBLESHOOTING! The mark of a neophyte servicer is the use of the "diddle stick" in troubleshooting. So-called alignment problems do not occur suddenly, but typically occur over a very long period. When alignment shifts suddenly, it isn't an "alignment problem" at all, but rather a component failure. Alignment causes a certain amount of wear and tear trauma, so constant tweeking is likely to add more faults than it may correct.

You might also want to keep a small supply of certain service chemicals on hand. Repair and maintenance jobs sometimes require cleaning of switches and potentiometers, so a spray can of cleaner (*Blue Stuff,* for example) is handy. You'll also need a small tube of either silicone grease or white grease (*Lubriplate* is a popular brand), some heat transfer grease for power transistors, and a freeze spray for locating intermittents. For that really professional touch on equipment with PCBs, buy a bottle of PCB cleaner designed to remove solder resins. Cleaning cabinets can be done with special cleaners, but they're expensive, so I always use modest amounts of soapy water on soft sponges or cloths.

I also recommend that you order the shop manual for your rig. Most rigs come with a manual that's fine for operating (it may even contain a schematic diagram), but not much good for troubleshooting. The shop manual must be purchased separately. Some rigs include a postcard or order form with the rig's documentation to make ordering a shop manual easier. In other cases, you'll have to write to the manufacturer or importer and ask about the availability of the manual. You should also consider buying service aids such as PCB card extenders, which allow you to operate PCBs out of the rig for troubleshooting, as well as patch cords and other items the company offers to professional servicers.

conclusion

Repairing commercially made electronic equipment is not always a difficult matter better left to professional repair services. The Amateur with enough knowledge to pass a General or Advanced class license examination should be able to learn the job of trouble-shooting and repair — even if under a friendly "Elmer" the first time.

ham radio

a detailed look at probes

From simple to exotic, these devices invariably affect circuits

Although it's admittedly difficult to get excited about probes, this article may offer insight into, and perhaps even respect for, these devices. Despite their uncomplicated appearance, they're really much more than "just a piece of wire with a handle."

probes load down circuits

Most low- to medium-frequency oscilloscopes have a 1-Megohm impedance shunted by 8 to 50 pF of capacitance. Some scopes above 200 MHz have a 50-ohm or both a 1 Megohm and 50-ohm input impedance that is switch-selectable. In either case, when a scope's probe is attached to the circuit under test, this impedance loads the circuit and may alter its proper operation or even cause it not to work.

Figure 1 shows an attenuator probe and scope input. An attenuator probe contains a built-in voltage divider. Most typical values of these types of probes are 10:1 and sometimes even 100:1.

How does the probe affect the input signal? First, consider what happens to square wave and sinewave inputs. (With sinewave inputs we're especially concerned with amplitude and phase distortion.) Referring again to **fig. 1**, note that the probe and scope input essentially form an RC divider. Since R1C1 must equal R2C2 for equal attenuation at all frequencies, as R1 increases C1 must decrease. The capacitance at the probe tip can be reduced by going to higher values of attenuation.

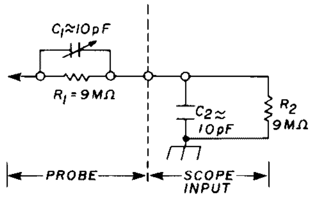


fig. 1. Illustration of a probe and scope's input circuitry.

Note: All figures courtesy Tektronix, Inc., unless otherwise noted.

measuring pulsed signals

Referring to **fig. 2A**, note that if $R_s = 200$ ohms and if $C_s = 20$ pF, then t_{r1} would be limited by the integration network of R_sC_s and would be equal to 2.2 $R_s \cdot C_s$ or 8.8 nanoseconds in this case. Using a typical passive probe such as the P6053B (**fig. 3**) with a 9.5 pF and 10 Megohm impedance then results in the circuit shown in **fig. 2B**. (R_p has been disregarded because it is much greater than R_s.) Looking at the risetime as 2.2 R_s ($C_s + C_p$) we obtain 13 nanoseconds. Mathematically

$$\frac{t_{r2} - t_{r1}}{t_{r1}} \times 100 = \frac{13\ ns - 8.8\ ns}{8.8\ ns}$$

$$\text{and } \frac{C_p}{C_s} \times 100 = \frac{9.5\ pF}{20\ pF} = 48 \text{ percent}$$

Next, let's look at what happens when we use a P6048 probe (**fig. 4**) with its 1 pF, 10 kilohm characteristic. Referring to **fig. 2C**, note that by using Thévenin's theorem that $t_{r3} = 2.2\ R_2\ (C_s + C_p) = 7.7$ nanoseconds now. This percentage change is dramatically less than that caused by the P6053B. Mathematically,

$$\frac{7.7\ ns - 8.8\ ns}{8.8\ ns} \times 100 = 12 \text{ percent change}$$

Interestingly, though, by not degrading the signal by slowing its risetime, the probe actually modified the source resistance to decrease the risetime, making it faster than it should be. This, however, was at the expense of output amplitude, which decreased to 83.3 percent. Recall that in the first example there was no change in signal source amplitude when the probe was applied.

By Vaughn D. Martin, 114 Lost Meadows, Cibolo, Texas 78108 and Billy W. Davis, 8914 Rich Quail, San Antonio, Texas 78251

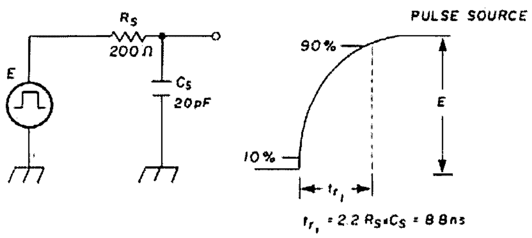


fig. 2A. Typical pulsed signal source.

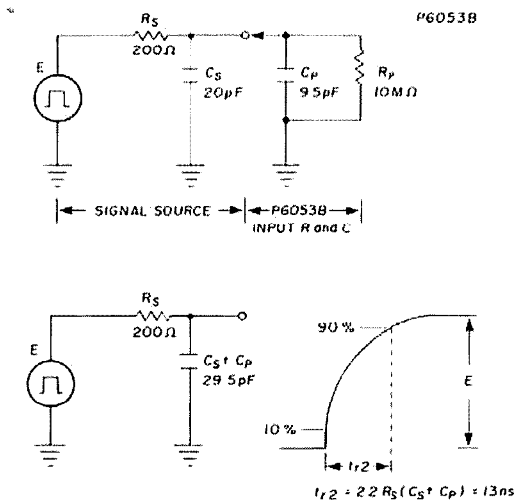


fig. 2B. A P6053B probe added to this typical signal source.

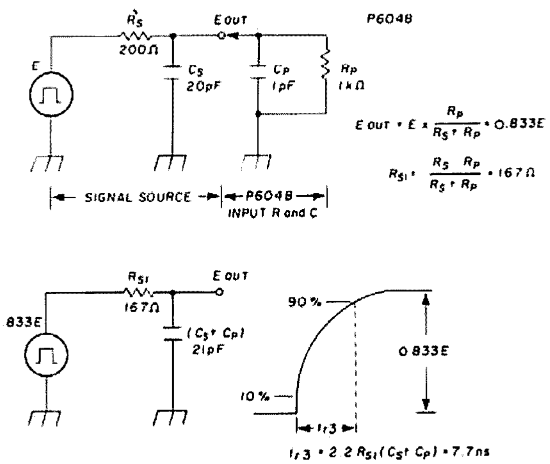


fig. 2C. A P6048 probe added to this typical signal source.

These two examples demonstrate that when measuring pulsed or square wave signals, low capacitance is desired when measuring risetime, but *high resistance* is more important when measuring amplitude. Also, selecting a low impedance test point in a circuit is best when measuring both risetime and amplitude.

Using the same probes and the same circuit, let's see how amplitude and phase relationships are affected when the signal source is a sinewave oscillator. Refer to **fig. 5** and note that we now have a 10 MHz source; at these higher frequencies, the X_p and R_p of the P6053B probe change (**see fig. 6**). Applying the P6053B probe to the source, the output voltage drops to 94 percent of the generator voltage. This represents a 3 percent drop from the 97 percent output voltage normally obtained from the source.

Applying a P6048 probe to the same circuit causes an output voltage of 81 percent of the generator's open circuit voltage and 16 percent from the unloaded voltage condition (**fig. 7**).

Because all probes contain a capacitive element, phase relationships or shifts will naturally occur. Refer to **fig. 8** and consider an amplifier driven from a 10 MHz, 50 ohm source and having an output impedance

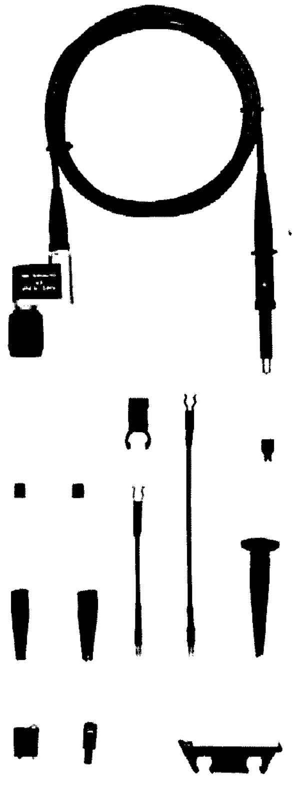


fig. 3. P6053B high-impedance probe.

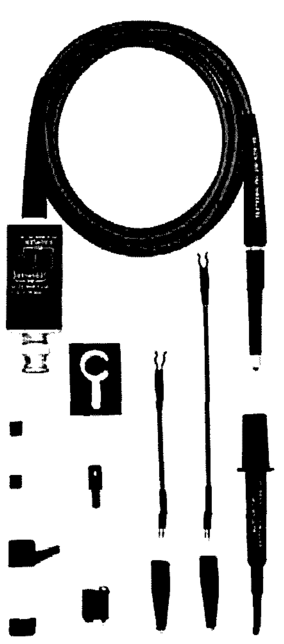


fig. 4. P6048 low-capacitance probe.

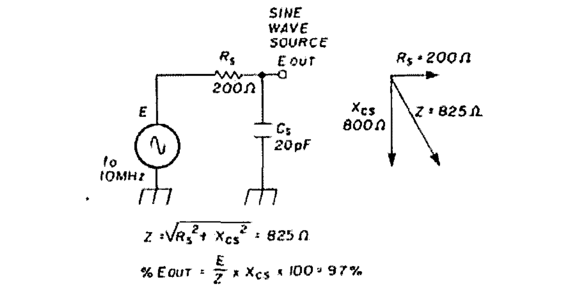


fig. 5. A typical sine wave signal source.

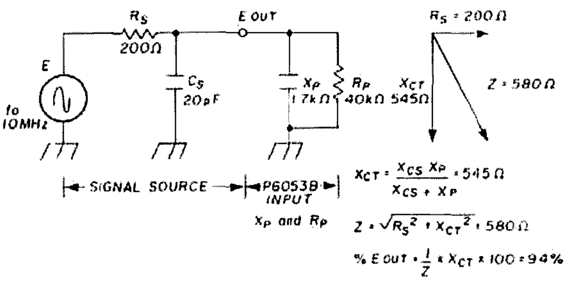


fig. 6. A P6053B probe applied to this typical source.

of 2 kilohms. Examining the input and output using two 10 Megohm 10 pF probes graphically shows a 49 degree phase shift difference in the points in the circuit being examined.

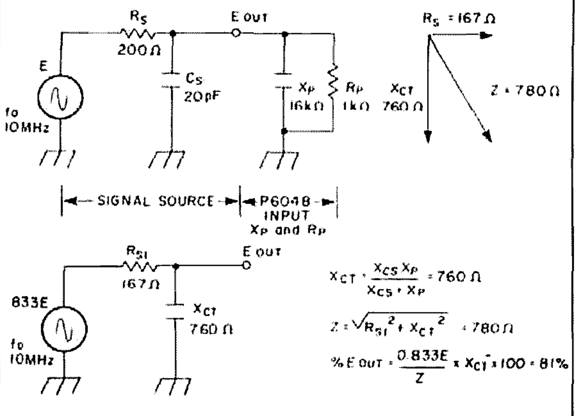


fig. 7. A P6048 probe applied to this typical source.

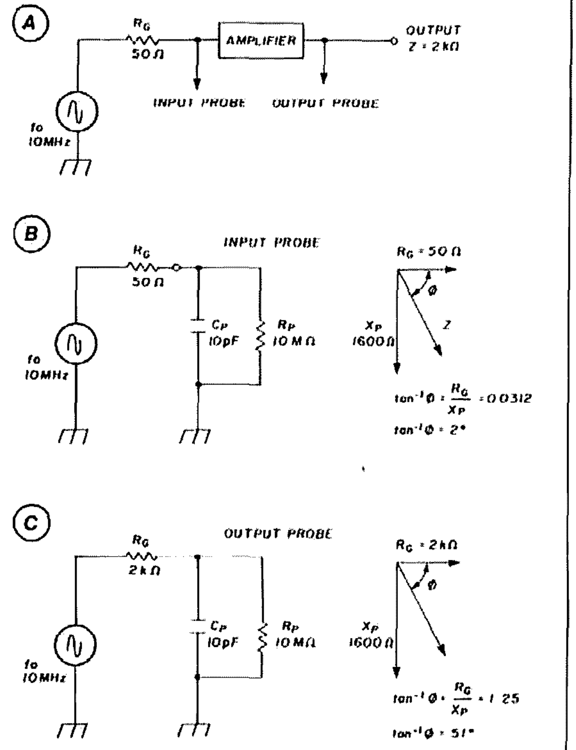


fig. 8. (A) Typical amplifier circuit with differing input and output impedances. (B) Phase shift caused by applying P6053B probe to the amplifier input. (C) Phase shift caused by applying P6053B to amplifier's output.

Next, refer to fig. 9. Note that with 1 kilohm, 1 pF probes the phase difference is now only 2 percent. But the price paid for this is an attenuation in signal — it's only 67 percent of the original amplitude.

application determines probe needed

Here are some general guidelines to follow when making measurements with a passive scope probe.

- Always check the probe compensation on the oscilloscope being used to make the measurement.
- Choose the lowest *impedance* test point possible to view the signal.
- When making risetime measurements:

  Choose a probe with R and C values as low as possible.

  Scope and probe risetime should be short relative to the signal risetime.

  Observed risetime is approximately equal to the square root of the sum of the squares of all risetimes in the system. These risetimes include that contributed by the signal source, the probe and the scope.
- When making amplitude measurements:

  For sine wave measurements, choose a probe that has the highest input impedance at the frequency of interest. Remember, loading error changes with frequency.

  For pulse measurements, choose a probe that has a large input resistance relative to the source impedance. Input C is of no concern if pulse duration is about five times longer than the input RC.

active probes

Two prime advantages of active probes include the following: isolation (provided between the measurement point and the probe cable and scope, allowing high input resistance and low capacitance to be achieved) and full bandwidth (obtainable without input signal attenuation).

Most active probes are compatible with either 1 Megohm or 50 ohm scope inputs without using external adapters. When working in the 50 ohm mode, a 50 ohm cable can be used to extend the probe length without increasing capacitive loading. However, longer cables will slow the risetime.

A typical active probe such as the Tektronix P6201 in **fig. 10** has a probe bandwidth of from DC to 900 MHz with a risetime of 0.4 nanosecond.

To show how much better an active probe is than a passive probe, let's go back to the same circuits we've been comparing and see what happens to the circuit once the active probe is connected to it (see **fig. 11**). Realizing that the P6201 has a 1.5 pF capacitance, its loading effect is only 8 percent, increasing the pulse risetime from 8.8 ns to 9.5 ns or mathematically:

$$\frac{t_{r3} - t_{r1}}{t_{r1}} \times 100 = \frac{9.5\text{-}8.8}{8.8} = 8 \text{ percent}$$

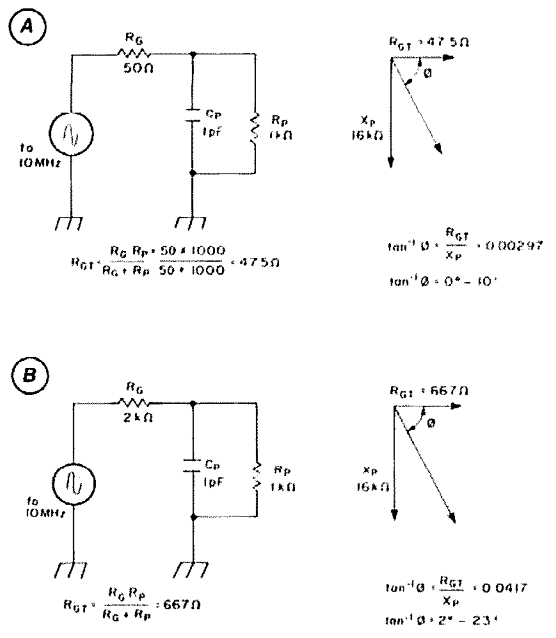


fig. 9. Phase shift caused by applying a P6048 probe to an amplifier's (A) input and (B) output.

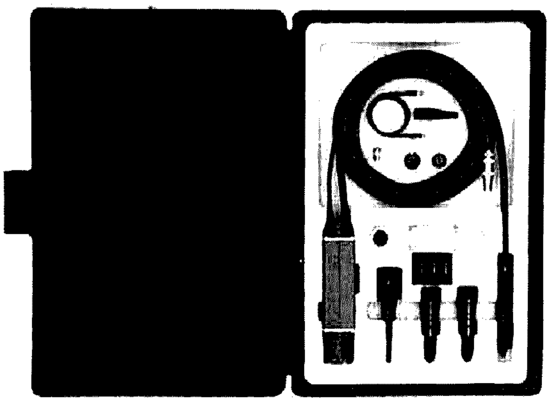


fig. 10. A typical active probe, the Tektronix P6201.

This 8 percent value is a marked improvement over the 48 percent increase in risetime caused by a 10x high impedance *passive* probe like the P6048.

measuring low-level signals

One prime advantage of an active probe is full bandwidth at 1x attenuation with minimum circuit loading. **Figure 12** shows that the active probe increased the risetime from 8.8 ns to 10 ns for a 14 percent change. Although somewhat greater than the 8 percent change

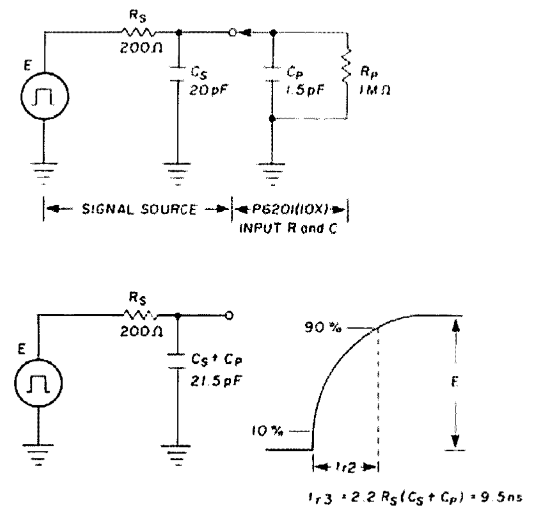


fig. 11. A P6201 (10X) active probe added to a typical pulse source.

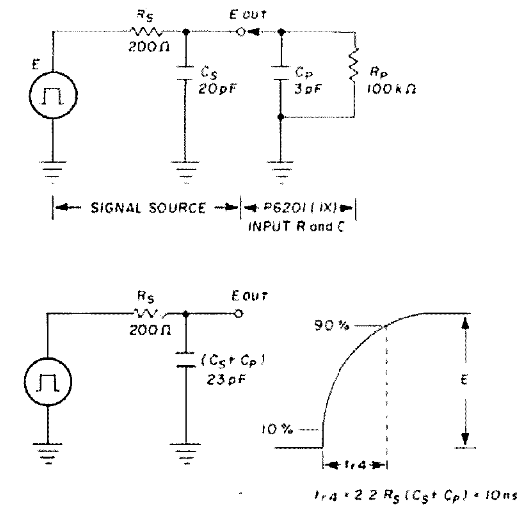


fig. 12. A P6201 (1X) active probe added to a typical pulse source.

of the P6201 (10x), it virtually had no effect at all on the signal amplitude. But in fairness to the passive probes, it must be noted that at lower values of source impedance or with slower risetimes the small difference in measurement error may not justify the difference in cost between the active and passive probes. But in the final selection process it might be advisable to observe the following general guidelines:

• Full bandwidth is provided with no signal attenuation using the 1X configuration.

• The active nature of the probe provides the high input impedance characteristics of most passive probes and the low input capacitance of passive probes designed to work into 50 ohm inputs. These features yield the best features of both — minimum risetime and minimum pulse-amplitude error.

• Impedance selection to permit use with either 50 ohm or 1 Megohm inputs is usually provided.

• Probe length can be extended through the use of a 50 ohm cable without increasing probe loading.

• Over-voltage capability is typically provided. However, to minimize the likelihood of over-voltage, the highest attenuation configuration should always be used when probing unknown voltages.

• Dynamic signal range of the active probe is not as great as that of a passive probe. For example, the P6201 (1X) can handle signals up to +600 mV. This can be extended to ±60 volts using the 100X attenuator. DC offset provides a measurement window of ±5.6 volts using the probe alone, with the range extended to ±200 volts using the 100X attenuator.

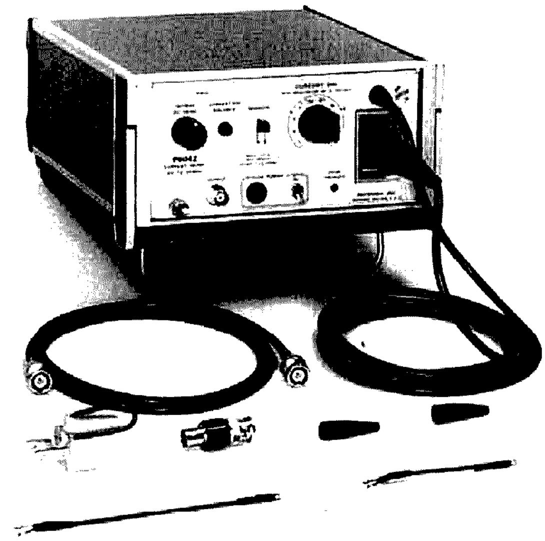


fig. 12A. The P6042 current probe and accessories.

the current probe

Now let's turn our attention to a measurement tool often overlooked — the current probe. Current probe measurements are particularly applicable for high impedance measurement points where the voltage probe would significantly alter the circuit characteristics.

The current probe offers the lowest circuit-loading of any available probe. There is, however, an insertion impedance reflected into the circuit under test, which consists of a series resistance shunted by a small

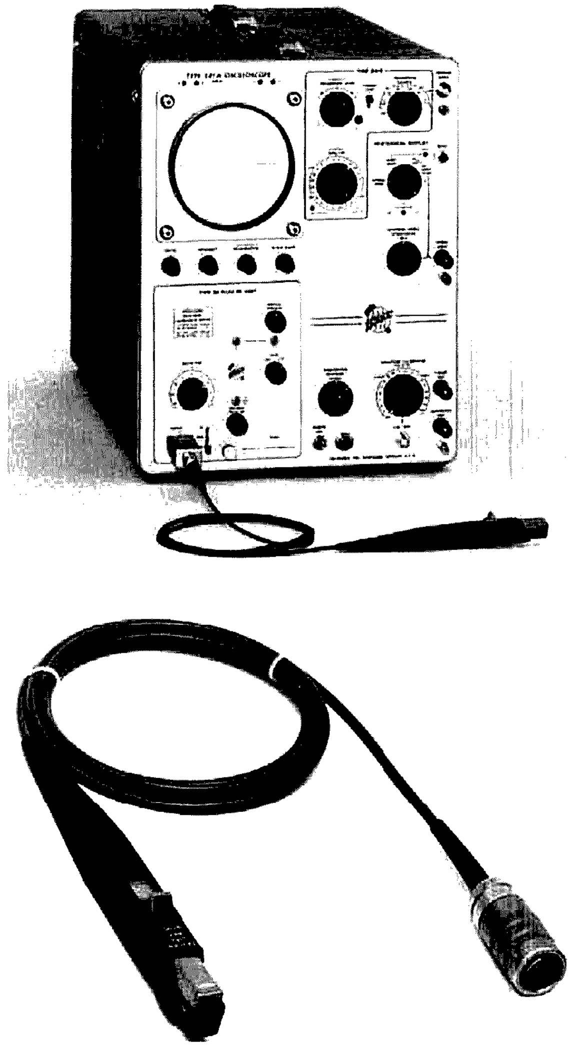


fig. 13A. Examples of closed-core head current probes.

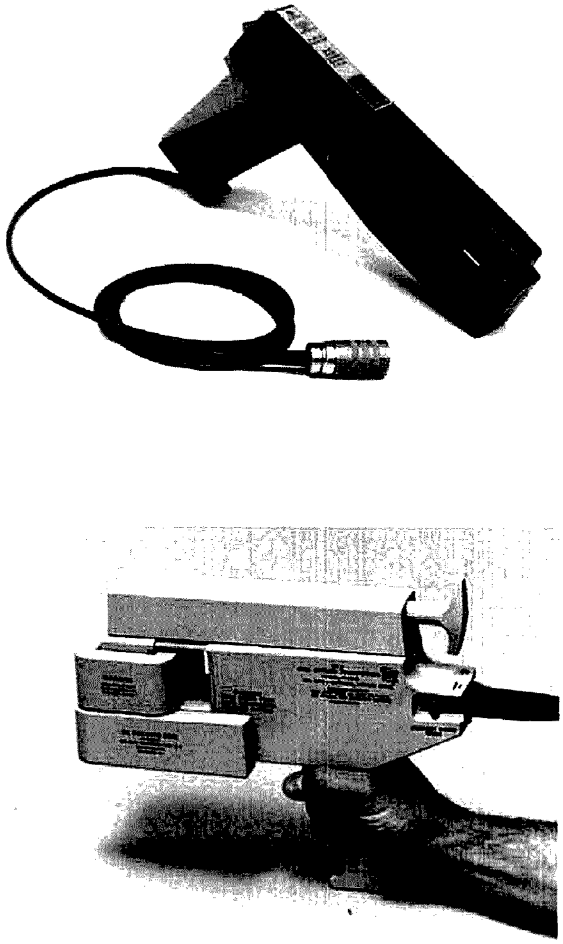


fig. 13B. Examples of split-core head current probes.

inductance which is associated with the inserted impedance of the current-sensing unit in the probe head. The P6042 current sensing probe is a DC to 50 MHz device with insertion impedance of 0.1 ohms at 5 MHz, (**fig. 12A**). Therefore, to realize an amplitude measurement error of no greater than 2 percent the signal source impedance should be 50 times the insertion impedance or 5 ohms in this case. But let's take a closer look at this head. It uses a Hall Effect* device to yield tilt-free display of the current waveform. This also supplies the DC and low frequency information to the P6042 amplifier. In this amplifier, this low frequency information is combined with the high frequency component of the signal to yield the output.

Another consideration in using current probes is the capacitive loading from the probe to the circuit. The coupling — the only shunt loading placed on the circuit by the probe — will vary according to the size and type of wire or current conductor. For example, while a No. 20 AWG wire will show approximately 0.6 pF, a No. 14 wire will have approximately 1.5 pF. The majority of capacitance occurs between the shielding and the current sensing unit. This can be minimized by using the probe ground lead when working with large voltage swings of high-frequency signals.

There are two types of current probes (**figs. 13A** and **13B**): the *closed-core unit*, in which the wire is

*The Hall Effect describes the phenomenon of what happens when a conductor through which a current is flowing is placed in a magnetic field: a difference in potential (the Hall voltage) is generated between the two opposite edges of the conductor in the direction perpendicular to both the field and the current.

threaded through the head and the *split-core unit*, which provides for a portion of the core to slide back, allowing the current-carrying lead to be inserted without breaking the circuit.

A different set of terms is used to describe the operational characteristics of a current probe than is used with a voltage probe. The most important parameter is the Amp-Second Product, which is directly related to the flux saturation of the transformer core. Effectively, the Coulomb* charge (one volt is the potential difference between two points in an electric circuit when the energy involved in moving one coulomb of electric charge from one point to the other is one joule) under one pulse is integrated to determine whether it will place the current transformer into saturation.

The following are general considerations related to current probe use:

- The current probe can be considered complementary to the voltage probe in that while the voltage probe requires low impedance points for accurate measurements, the current probe requires higher impedance points.
- The current probe exhibits lower loading than any voltage probe. This generally implies minimum signal amplitude attenuation and minimum risetime inaccuracies.
- Where information on current supply requirements is needed, primarily current into capacitive elements, the current probe is almost a necessity.

a more exotic probe

The last type of probe to be considered is the sophisticated, cleverly designed HP5363B time interval probe (**fig. 14**).

An electronic counter works on a simple principle in the time interval mode. First, any frequency counter has a main gate that when opened, allows pulses from an extremely accurate quartz crystal oscillator to be counted and accumulated in a counting register. The main gate remains open and the register keeps counting the crystal's frequency as long as a measurement is being made. In the time interval mode, there is a finite elapsed time between two events. These two events, the START and STOP events, take the form of the two independent inputs on a frequency counter. The problem arises when the two channels (the START and STOP inputs) have unequal delays through their probes, cabling, and input circuitry. Also, a counter's input circuitry has been designed for optimum performance at detecting zero crossings. This makes the measurement of risetimes, propagation delays and slow rates very difficult because its limited

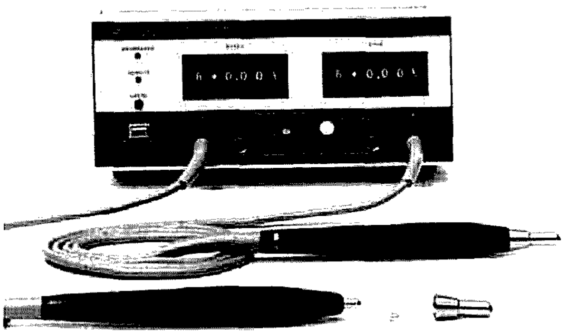


fig. 14. The HP5363B time-interval probe. ***(Photo courtesy Hewlett-Packard.)***

trigger range is typically ± 1 volt or less. A slight uncertainty or ambiguity in the exact triggering point results.

The HP5363B time interval probe can be very useful in this situation. The user goes through a procedure that basically involves grounding the probe to be calibrated and pressing a front panel switch. This action causes the reference voltage, V_r, to move up or down in a stair-step fashion in very precise 1 mV steps until the device triggers. Knowing the exact value of V_R helps and the dynamic range of this probe is $+9.99$ volts to -9.99 volts and is presettable in 1 mV steps by the thumbwheel switches on the front panel. This probe eliminates the need for attenuators and allows more accurate measurements nearer the top and bottom of waveforms than would be possible without this device. The probe also has a pullable engaging type switch that adds 10 nanoseconds to the measurement to compensate for less than ideally matched input channels. Naturally, though, this 10 nanoseconds has to be added to the final reading if you engage this feature on the time interval probe.

conclusion

The judicious selection of a probe results in more accurately derived readings, and this enhanced accuracy will be evident in both more precise amplitude and phase measurements with a passive scope probe and more accurate current measurements with an active current probe. And if you have to make very precise time interval measurements (stop-to-start intervals, for example,) between events using a frequency counter, you now know of an accessory for the frequency counter's probes that minimizes and nulls out the differences in each channel's probe and its input circuitry.

ham radio

*A Coulomb is mathematically equal to $\frac{W}{V}$ or one joule divided by one volt.

ham radio TECHNIQUES

Bill Orr W6SAI

"I have seen the future — and it works"*

It may be that Amateur Radio is undergoing a technical revolution as great as those revolutions brought about by the switch to single sideband and to VHF repeaters. Both developments changed the face of Amateur Radio within a few short years. Amateurs of the 1940s would be astounded if they were magically transported through time to visit a modern Amateur Radio station.

The microcomputer is leaving its mark on Amateur Radio today. Last winter, for example, the first 20-meter SSB computer controlled contacts were made via a VHF/UHF repeater and remote link (**fig. 1**). The heart of this unique communication circuit was the newly-developed Shackmaster® station control unit (**fig. 2**) developed by Ed Ingber, WA6AXX, and his associates at Advanced Computer Controls, Inc., of Cupertino, California.

The Shackmaster is a station accessory that permits remote control of a home station via digital command given over the air or via a telephone line. In the case I'm talking about, N6IPE (Don Melchoir), controlled his 14 MHz SSB station via a remote VHF radio link. And this is *complete* — receiver and transmitter band scanning, mode changing, band changing and on/off control — just as if the operator were sitting at the station console!

*John M. Reed, 1920.

crossband linking

Imagine yourself driving along the highway. Your vehicle is equipped with two FM transceivers, one on the 440-MHz band and the other on the 1296-MHz band. At the moment you're listening on 443.875 MHz, which is the output frequency of the hill-mounted repeater. The down-link is 443.875 MHz, and an FM receiver at the 20-meter home station is listening to

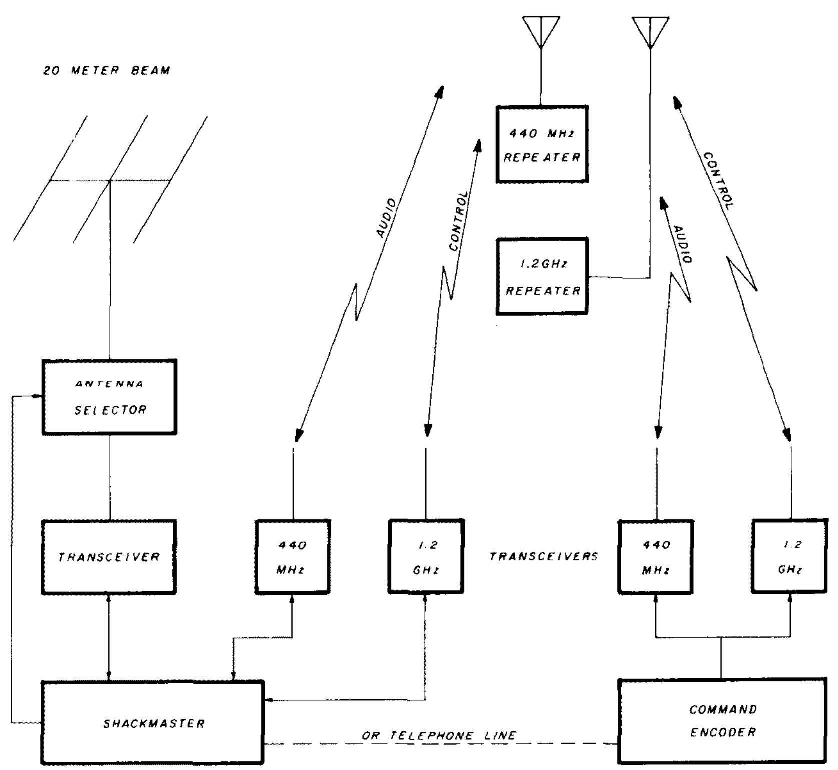


fig. 1. The Shackmaster supports voice, frequency and mode control of 20-meter transmitter via two UHF links through repeaters. In the N6IPE experiments an ICOM-751 computer-controlled HF transceiver was used. Instructions are loaded into memory circuits of transceiver. DX to date includes Guam, Kwajalein, and Venezuela.

fig. 2(A). Shackmaster 100 Station Controller* allows operator to control home station over the air or over the phone lines. Unit ties together home equipment and allows deposit of electronic messages in its "mailbox."

this channel. This link provides the two-way voice channel. The 1296-MHz channel is provided for frequency control of the home station and for transmit-receive control. Everything's ready for operation.

Via the 1296-MHz link, you key the command "HF LISTEN" by sending your individual code number plus *74*. You now key in *3* and the "SCAN UP SLOW" command starts the transceiver to slowly tune up the band. To "STOP SCAN," you key in *2* and to scan down, you key in *1*. Other commands allow fast scan up or down, and frequency jumps of 20, 100, and 500 Hz. The keyed command *5* allows you to shift to the auxiliary VFO, which can be programmed separately.

Now you've tuned in the signal you wish to call. As soon as the opportunity presents itself, you pick up the 440 MHz mic, key in **75* and call. To return to listen, you key in *5*.

When the contact is finished, you

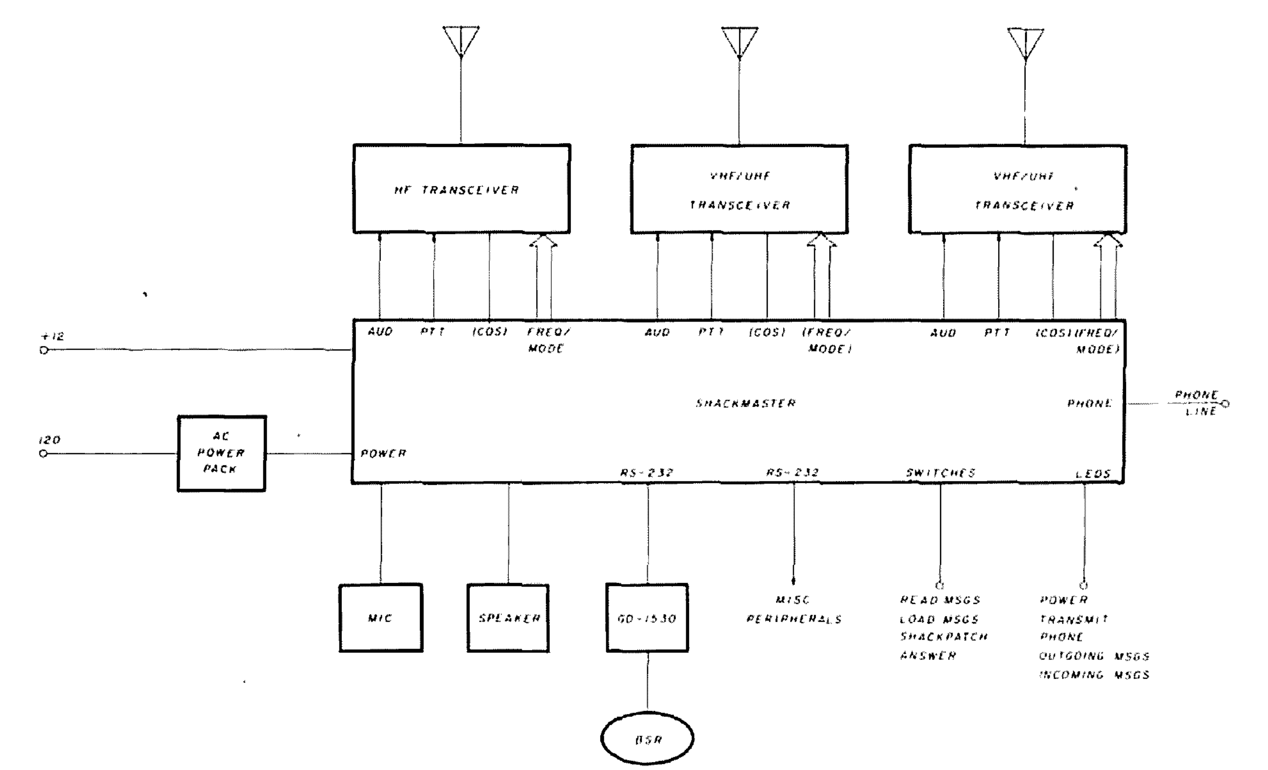
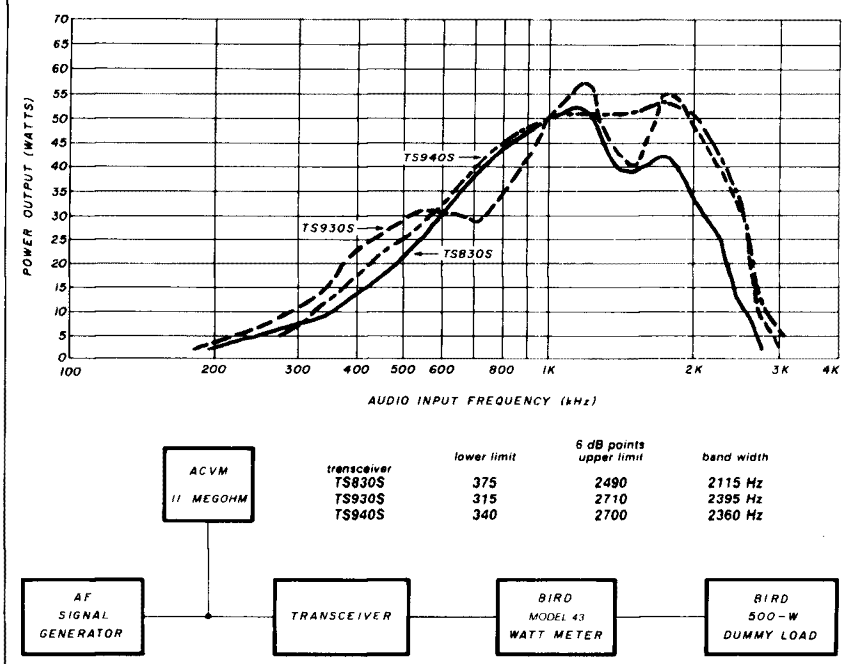


fig. 2(B). Block diagram of Shackmaster.



transceiver	lower limit	6 dB points upper limit	band width
TS830S	375	2490	2115 Hz
TS930S	315	2710	2395 Hz
TS940S	340	2700	2360 Hz

fig. 3. Test setup for measuring audio response of transceiver.

decide you want to see what's coming in from the west. A beam heading of 270 degrees will do the job. You key in **4* to actuate the rotator, followed by *270*. Your antenna is now aimed west.

All this may sound very complex and confusing at first, but it quickly becomes second nature and provides an enormous amount of control capability with a touchtone pad.

The home station makes use of a computer-controlled transceiver, broadband linear amplifier and computer-controlled antenna tuner. In the case of N6IPE, the transceiver was an ICOM-751, the linear amplifier was an ICOM 2KL solid state, broadband device, and the antenna tuner was an ICOM-AT500. The IC-751 was run in the memory mode and the various commands were loaded into the memory.

As you become experienced with the system, you can query the Digital Voice Synthesizer, which will tell you the frequency the 20-meter equipment is tuned to. The capabilities are infinite. Too much QRM on 20 meters? Key *72400* on the touchtone pad and you've instantly QSY'd to 7.240 MHz.

And finally, when you get to your destination and leave your mobile rig, you can still activate your home station via the telephone line!

Well, it all sounds like lots of fun. No doubt you'll be hearing more and more remote-controlled, computer operated stations on the HF and VHF bands during the coming months. The day when two computers talk to each other is not far away in the amazing world of Amateur Radio.

audio response revisited

In my April column I discussed some of the variations in the audio response curve of some representative SSB transceivers. The conclusion was that the passband of some of them left much to be desired. Recently Tiff, W6GNX, had the opportunity to run an audio passband check on three more HF transceivers: the Kenwood TS-830S, the TS-930S, and the new TS-940S. The test was accomplished by injecting an audio tone into the "phone patch" port and measuring the power

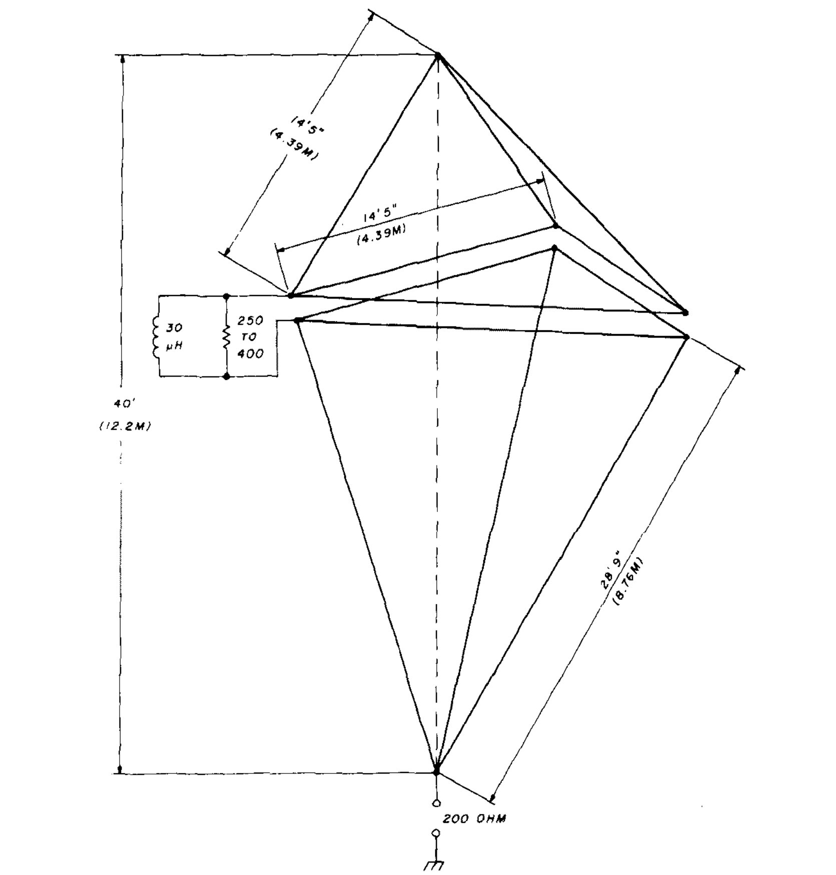


fig. 4. The modified Australian monopole covers 3.5-17.5 MHz. Original design used 377-ohm resistor in RL network.

output of the transceiver as the tone was swept across the audio passband. Sufficient instrumentation was used so that the tone level was constant and the transceiver always operated in the linear mode. The results are shown in **fig. 3**.

The TS-830S and TS-930S both exhibit the characteristic "bumpy" audio response common to that derived from a multi-pole filter that controls the passband. (Compare the curves with that run on the FT-980, as shown in the April column).

Now observe the passband curve of the TS-940S; it's nearly devoid of the filter "humps" and shows a nearly flat frequency response from about 700 Hz to 2300 Hz (less than 1 dB variation). Of interest, too, is the enhanced high-frequency response of both the TS-930S and TS-940S, as compared to the TS-830S.

On-the-air tests of all three transceivers over a period of time showed that it was easy to distinguish the TS-830S from the other two transceivers. It just didn't sound as clear and penetrating, even though the same microphone was used for all tests.

It was even more difficult to differentiate the TS-930S from the TS-940S. I could do it because I knew what Tiff's voice sounded like in person. But the on-the-air difference was marginal.

In any event, the audio passband of the TS-940S is a step in the right direction. As AG6K said, "Perhaps we can now stop sounding like Donald Duck on the air!"

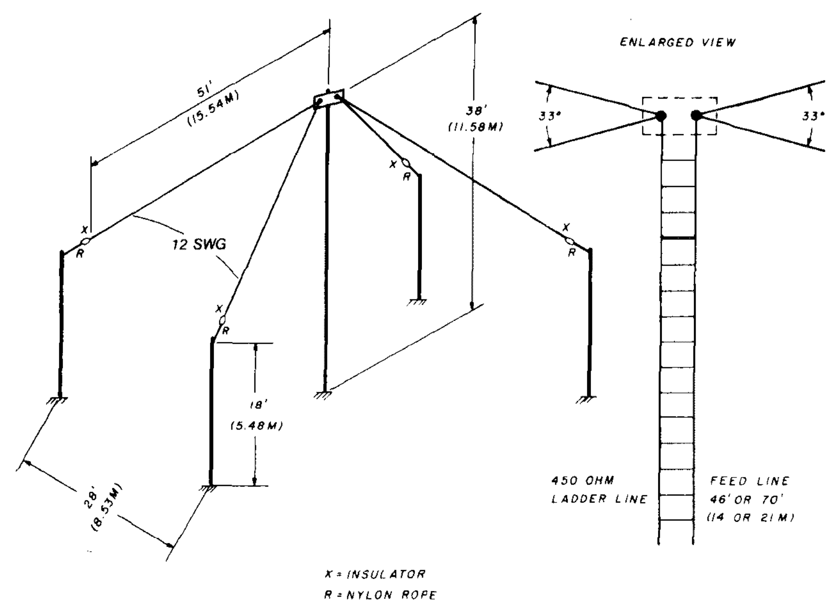


fig. 5. "All-band" antenna patterned after G5RV design. Angles between wires is not critical.

the "Australian monopole"

In my April, 1984, column I described a wideband monopole antenna. Recently I received a letter from Tom, KA2APX, who has built and used it. He says, "I've had several hundred hours operating time on the antenna and am very pleased with the results. I've seen a received signal increase between 12 to 15 dB, as measured on an SP-600 S-meter as compared to an inverted-V at resonance. Incoming signal strength has been good overall, with VK and ZL signals running reasonably well above the noise floor on 75 and 160 meter SSB."

Tom is using a 100-watt, 250-ohm resistor in the antenna network and a 60 inch diameter cage instead of a 72 inch one. He concludes, "Many stations, whom I regularly contact, have indicated an interest in the antenna because of its compact length, radiation efficiency, and bandswitching ease."

Recently *Amateur Radio*, the publication of the *Wireless Institute of Australia* described a new version of this interesting wideband antenna. The description was written by Ron, VK3AFW; the general layout of the antenna is shown in **fig. 4**. It's a kite-shaped, three-dimensional affair, about 40-feet high, that could be suspended from a single pole with crossarms to support the widest part of the structure.

The upper and lower portions of the antenna are connected with a simple R-L network. Experience has shown that the resistor should be noninductive and have a power rating equal to about 10 percent of the power output of the transmitter.

The feed point resistance of the antenna is about 200 ohms, so a 4:1 balun is used to match a 50-ohm transmission line. The frequency range of the antenna is 5:1, so this design should presumably cover the range of 3.5 MHz to 17.5 MHz. Ron has found that bumps in the SWR response curve can be moved about or eliminated by varying the value of the shunt resistor in the antenna network.

As with any vertical antenna, a good ground system is required, although VK3AFW reports good results when using only a single 6-foot ground rod.

The whole family of so-called "Australian" antennas (dipole and monopoles) was covered in a technical paper, "Low Profile Radiator for HF Surface and Skywaves," by R.R. Treharne, published in the *IRRECON Digest* (Melbourne, Australia, August, 1981).

It would seem to me that an antenna of this type — one that could be packaged like an umbrella, capable of being opened and erected by one person — would make an excellent portable antenna for military or commercial use, and represent an interesting challenge in experimental design for Radio Amateurs.

a simple "all-band" antenna

In my February column I discussed the so-called G5RV multiband antenna, which is very popular in Europe but not as well-known in the States. In brief, it's a 102-foot (31.09 meter) wire, center-fed with tuned feeders and a Transmatch,® or other form of antenna tuner.

I understand many of the dedicated QRP stations use this antenna and the design has recently resurfaced in the fine QRP Column hosted by George Dobbs, G3RJV, in *Radio Communication*, the flagship publication of the *Radio Society of Great Britain*. The basic design is shown in **fig. 5**. It consists of two G5RV antennas, 33 degrees apart, fed in parallel by an open-wire transmission line. It's an ideal antenna to mount above a single-family residence, with the apex supported on a pole strapped to the chimney.

Both 300-ohm ribbon line and 240-ohm oval line have been used with this interesting antenna (taking into account the propagation factor of the line). A center height of 38 to 45 feet (11.58 to 13.71 meters) is recommended and end heights of 14 to 18 feet (4.26 to 5.48 meters) are satisfactory. And, as G3RJV says, "The antenna is cheap, simple and can, because of its inverted-V nature, fit into a surprisingly small space." That's not bad for an antenna that covers 3.5 to 28 MHz and exhibits some power gain on the higher frequency bands.

ham radio

This unit features azimuth and elevation memory storage, manual or clock control, digital readout and a 24-hour clock

a digitally-controlled satellite tracker

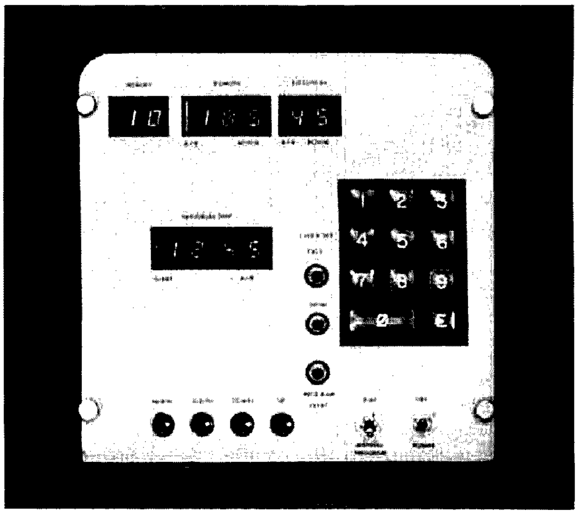


Photo A. Controller mounted in a BUD cabinet. The RUN-PROGRAM/MANUAL toggle swich shown here has since been changed to a 3-position rotary switch.

For satellite tracking *two* rotators are required for pointing the antennas in both azimuth and elevation. But constant adjustment as the satellite moves across the sky — which can be especially troublesome when you're trying to track the fast-moving, low orbit, Russian satellites — is, in fact, not necessary at all.

This system provides 1-degree tracking resolution. The unit (**Photo A**) includes a 24-hour clock and a memory for azimuth, elevation and time data. A small keyboard loads data for storage so that hands-off tracking can be accomplished by loading data before starting operations. Manual control is provided for positioning the antennas for start-up or manual tracking if desired.

the rotator

The most common rotator available is the U-100 TV rotator fròm Alliance. It has a double-field coil for forward and reverse, coupled through gears to the mast shaft. Every ten cycles of the 60-Hz power line turns the mast shaft 1 degree. A disc in the gear train mechanically closes a set of contacts every 10 degrees of rotation; this signal steps the control box. While 10-degree increments are a bit coarse for satellite tracking, the unit is inexpensive and readily available. A minor modification must be made to the rotator gear train, as explained later.

To get the rotators to turn with 1-degree resolution I use a counter clocked by the 60-Hz power line, which counts up or down as the rotator turns. Due to start up drag and antenna load, parity between mechanical position and readout cannot be maintained, and the 10-degree rotator signal is used to update the counter. Therefore, should the counter count ahead or behind, it is mechanically corrected by presetting the units counter to 5. For instance, when the counter has reached 14, mechanically the rotator is at 15 degrees. The rotator signal presets the counter to 15. Continuing on, the correction occurs again at 25 degrees, or every 10 degrees.

operating controls

The keyboard is used to load the memory with azimuth, elevation, and time data. The button in the lower right corner of the keyboard is the ENTER key and enters the data into memory. The memory program counter, clock and antenna aiming data are displayed. Azimuth is expressed in degrees from north; elevation in degrees from horizontal. Each seven-segment readout has an extra LED, which is normally used for the decimal point.

Some of these LEDs show internal functions. The A = B LED indicates that the data entered from the keyboard has successfully entered the memory. The START LED indicates the starting position for loading keyboard data.

Data is entered into the PROGRAM mode as follows: Depress the PROGRAM START pushbutton; the START LED lights momentarily. The first keyboard entry will load into the ten's hour bit of the clock. The next key will load into the unit's hour bit and so on across the program counter, AZ display and EL display. When all the data has been loaded, press the ENTER key to load data into memory. The A = B LEDs light to show successful memory loading. Press PROGRAM START again and we can load the next

By Rudolf E. Six, KA8OBL, 30725 Tennessee, Roseville, Michigan 48066

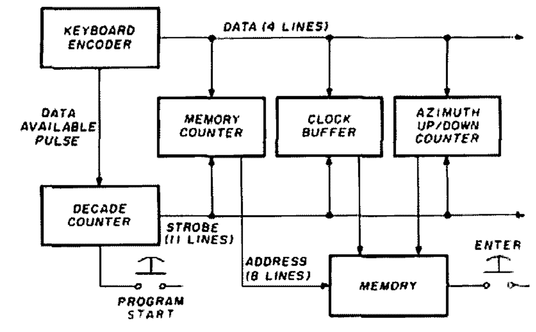
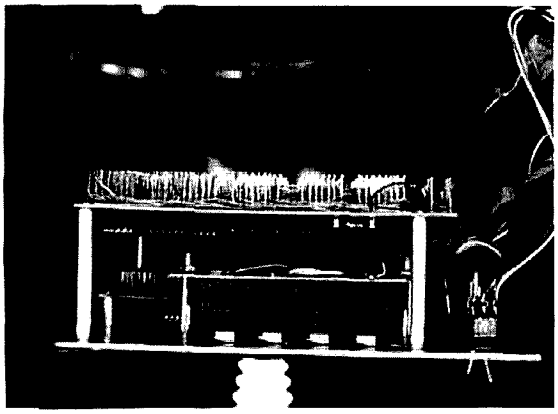


fig. 1. Diagram showing data logic when system is in PROGRAM mode.



Side view of front panel showing keypad mounting.

table 1. Typical data loading schedule for OSCAR 10 satellite. (December 21, 1984.)

universal *time*	*azimuth*	*elevation*
1747	267	10
1754	264	18
1800	261	26
1810	258	34
1823	252	43
1843	245	51
1944	230	57
2203	230	47
2316	235	38
0028	238	28
0141	238	19
0253	232	08

Programming the satellite tracker with antennas parked at 260-degrees azimuth and 80 degrees elevation:

	time	memory	azimuth	elevation	
PROGRAM	1747	01	260	80	ENTER
Program Start	1754	02	267	10	ENTER
Program Start	1800	03	264	18	ENTER
Program Start	1810	04	261	26	ENTER
Program Start	1823	05	258	34	ENTER
Program Start	1843	06	252	43	ENTER
Program Start	1944	07	245	51	ENTER
Program Start	2203	08	230	57	ENTER
Program Start	2316	09	230	47	ENTER
Program Start	0028	10	235	38	ENTER
Program Start	0141	11	238	28	ENTER
Program Start	0253	12	238	19	ENTER
Program Start	2600	13	232	08	ENTER
Program Start	0000	00	000	00	ENTER
RUN					

Note: The time value of 2600 at step 13 prevents turning by the rotators at the end of a run.

batch of data. Start with 01 in the memory counter and increment by 1 for a possible 99 steps.

Table 1 shows a typical loading schedule for OSCAR 10. The ROTOR LEDs show the closing of the cam switch in the rotator heads for every 10 degrees of rotation. They can be monitored for possible rotor jamming. In the RUN mode, time in memory is compared with the clock. When equal the program counter is incremented, and the rotators move to the new *azimuth and elevation data. In* MANUAL *mode the*

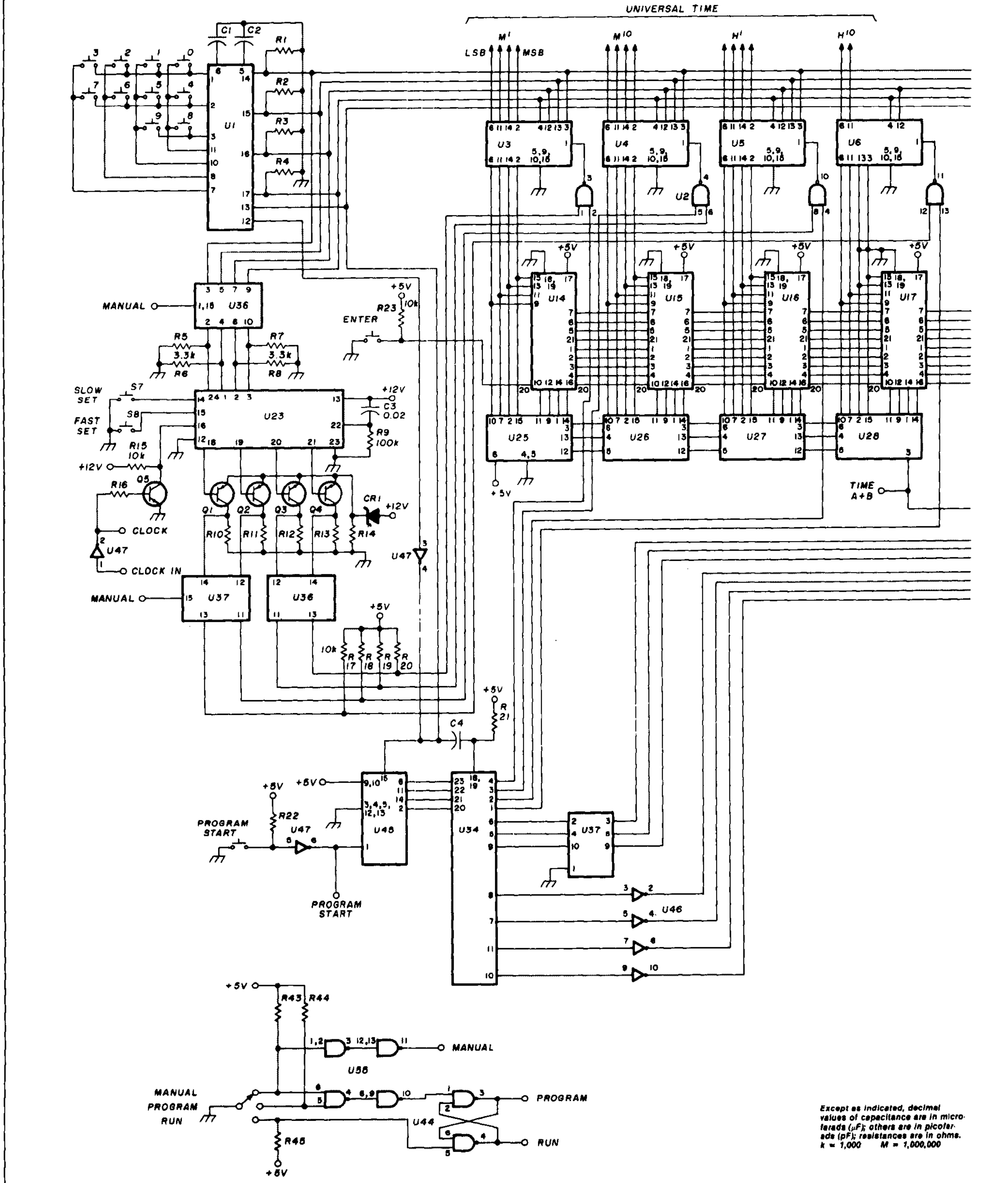
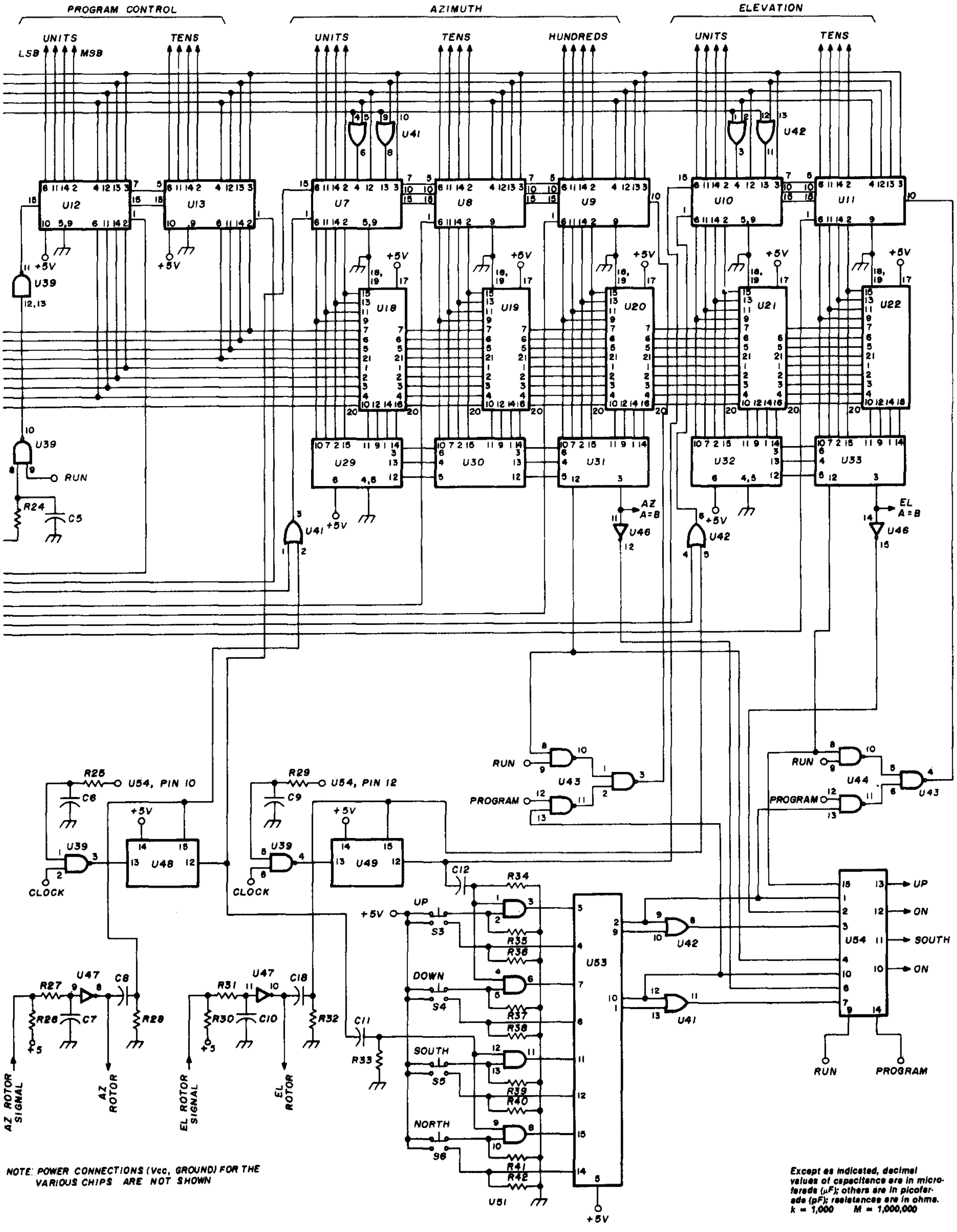


fig. 2 (part 1). Satellite tracker schematic diagram shows keyboard encoder, universal timer,



program counter, and azimuth/elevation counters.

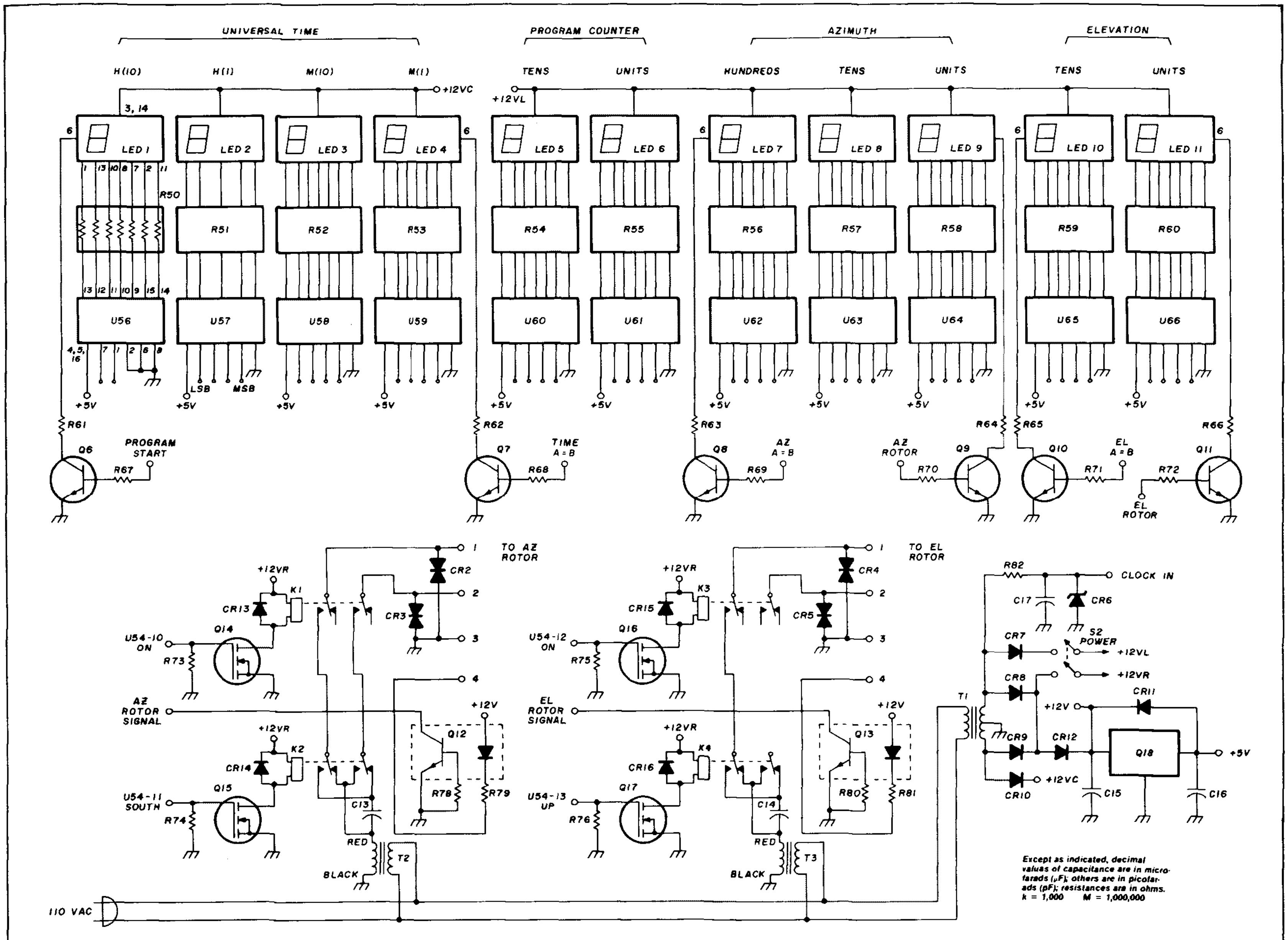


fig. 2 (part 2). Satellite tracker shows universal timer, program counter, azimuth and elevation display, and power supply.

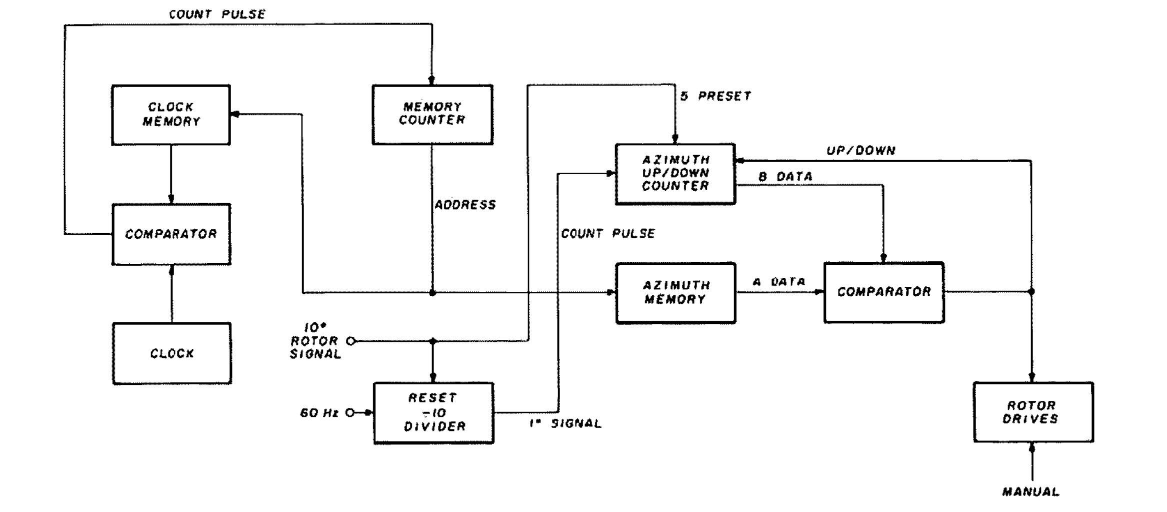


fig. 3. Data routine when in the RUN mode.

Photo B. Top view of front panel. Note keypad mounted on spacers.

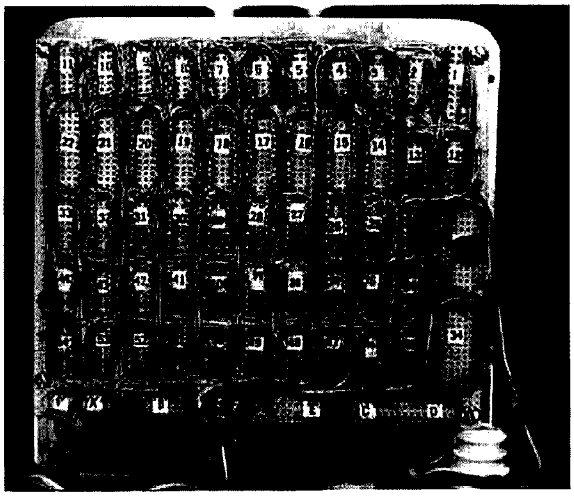


Photo C. Main circuit board showing wire-wrap. The numbers indicate ICs. Parts are mounted on DIP headers. Letters indicate sockets for interconnecting cables.

manual pushbuttons are available for positioning the rotators. A FAST and SLOW pushbutton is available to set the clock to universal (or Greenwich Mean Time) in the MANUAL mode.

basic operation

For simplicity the elevation circuitry will not be included. It is exactly the same as the azimuth circuitry.

• **Loading data from the keyboard (PROGRAM MODE), (fig. 1).** The key encoder (U1) encodes a keyboard array and outputs the binary value on four data lines (see **fig. 2**). This data presets the clock, AZ-EL and memory counters (U3 through U11). The key encoder also outputs a data-available pulse each time a keyboard entry is made. This pulse steps decade counter U45, U34. The output of the decade counter strobes each data counter sequentially until the data is loaded. We are now ready to store the information. The output from the memory address counter becomes the address for the memories (U14 through U22). The clock and AZ-EL counters provide the data input for the memories. Data is stored when the ENTER key is depressed.

• **Tracking the satellite automatically (RUN**

mode, fig. 3). During RUN mode the program counter is stepped by the output of the clock comparator (U25 through U28). When real time from the clock equals the time stored in memory the program counter increments to the next address. New AZ-EL data from memory (U18 through U22) is compared with the data in the AZ-EL counters (U7 through U11). Three signals are available from the AZ comparator (U29 through U31). If Data A = Data B, the data in both memory and counter is the same and the rotator will not move. If Data A is smaller than Data B, the rotator is turned on in reverse and the UP/DOWN counter is placed in the down mode.

A 1-degree pulse steps the counter until A = B and the rotator stops. If Data A is larger than Data B then the rotator is turned on forward, the UP/DOWN counter is placed in the up mode and again the 1-degree pulse steps the counter until A = B. The 1-degree signal comes from a decade divider (U48) which is clocked from the 60-Hz power line.

When the rotators are turning the counters are corrected every 10 degrees by a rotator strobe, which forces U7 to a count of 5 through OR gate U41 (pin 3). The decade divider is also reset by the rotator strobe from the Schmitt trigger U47 (pin 8).

• **Manual Mode.** In MANUAL mode the rotators are directly energized. The memory and comparator are not used. The manual pushbuttons each set a flip-flop (U53). Reset of each latch is synchronous with the zero count of the decade divider. This ensures manual operation in 1-degree steps. The select gate (U54) either routes MANUAL or RUN commands to the rotators. During the MANUAL mode the clock is also operational.

construction

A complete parts list is provided in **table 2**. All control circuitry was wirewrapped on the two boards. The board behind the front panel has the LED displays and drivers U56 through U66. The dropping resistors are DIP with seven resistors. The keyboard, purchased from Jameco, came mounted on a PC board, which was removed and the assembly remounted on a blank circuit board (**photo B**). The unit was then rewired to conform to the U1 keyboard decoder. Because the right row of switches is not used, the pushbuttons should be removed. (I have since used another keyboard sold by Herbach and Rademan, Inc. This keyboard doesn't have to be modified.)

The main circuit board contains all logic circuits in wirewrap sockets (**photo C**). Both boards are mounted to the front panel of a BUD cabinet with spacers. All interconnections were made with flat ribbon cable DIP plugs (**photos B, D**). Some care must be exercised in wiring the V+ and ground to the chips.

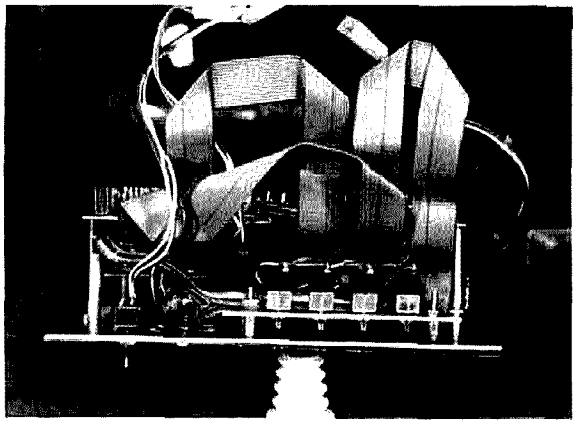


Photo D. Bottom view of front panel. Pushbuttons are mounted on a spacer. Note flat ribbon interconnect cables.

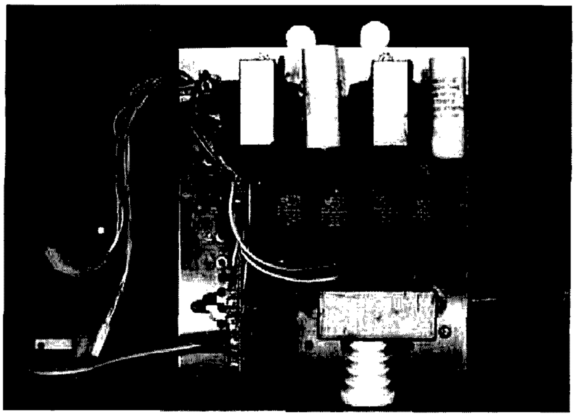


Photo E. Power supply chassis mounted in rear of BUD cabinet. Top transformers and capacitors are from original rotator drive. Small board to the left contains relay drivers, optical couplers and power supply. The +5 volt regulator is mounted on the chassis for cooling.

Each row was daisy-chained, and each chain was provided with a 0.01 μF bypass capacitor; all were then run to a common connection on the board. Two No. 18 wires with Molex plugs connect to the power supply. The power supply, rotator transformer, and control relay circuitry were mounted on the chassis back plate.

The rotators are very noisy when starting and stopping. Each output has a varistor (CR2 through CR5), which takes care of some of the noise, but good ground housekeeping is a must. Lead dress is especially important.

Coupling between rotators and circuitry is kept to a minimum by using relays and VMOS drivers for isolation (**photo E**). Optical couplers Q12 and Q13 are used for the rotator signal, and the 60-Hz clock signal is filtered through R82, C17 at the power transformer.

table 2. Parts list for the digital satellite tracker.

item	description
C1	10μF 10 volt tantalum
C2,7,10	0.1 μF 5 volt disc
C3	0.02 μF 5 volt disc
C4,6,8,9,11,12,18	0.01 μF disc
C5	1 μF 10 volt tantalum
C13,14	see text
C15	1000 μF 24 volt electrolytic
C16	50 μF 10 volt electrolytic
C17	0.2 μF 5 volt disc ceramic
CR1	3.3 volt zener
CR2 thru CR5	24 volt varistor GE V47ZA1
CR6	5 volt zener
CR7 thru CR16	1N4002
K1 thru K4	DPDT 12 volt coil
Q1 thru Q4	2N3905
Q5 thru Q11	2N3904
Q12,13	MCT-2 (4N25, LIT-1)
Q14 thru Q17	VN10KM (Radio Shack 276-2070)
All resistors are 10 kilohm, 1/4 watt except the following:	
R5,6,7,8	3.3 kilohm
R9	100 kilohm
R10,11,12,13,14	4.7 kilohm
R26,30	1 kilohm
R50 thru R60	470 DIP
R61 thru R66	470
R73 thru R76,78,80	68 kilohm
R79,81	620
R82	22 kilohm
S1	SP3T rotary miniature
S2	2PST toggle
S3 thru S6	SP2T momentary
S7,8,9*	SPST momentary N.O.
T1	24 VCT 2A (Stancor P-6377)
T2,3	see text
U1	74C922
U2,43,44,55	CD4011
U3 thru U13,45	CD4029
U14 thru U22	5101 RAM
U23	MM5312
U25 thru U33	74C85
U34	74C154
U36,37	74LS368
U39	CD4093
U41,42	74C32
U46	CD4049
U47	74C14
U48,49	CD4017
U51	74C08
U53	CD4043
U54	CD4019
U56 thru U66	74LS247
keyboard	Jameco K-19 TM23K222 (Herback & Rademan, Inc., 401 E. Erie Ave., Philadelphia, PA 19134)
displays	common anode, RHDP, LHDP (Hewlett-Packard HDSP-4030 or equivalent)

*"Slow-set" clock, "fast-set" clock, "program start."

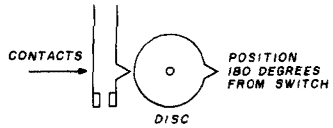


fig. 4. Repositioning the switch contact disc for 5-degree set in the Alliance rotator controller.

I have not experienced any problems with everything mounted in one unit, but keep in mind that CMOS is high impedance and will react to low-frequency noise.

Three separate pairs of wires supply +5 volts to the circuit boards. One pair is for the main logic card, one pair for the front panel LED circuitry and a final pair for the manual pushbuttons, mode switch and ENTER buttons. All three pairs originate at the regulator chip Q18. Since continuous power is needed for the clock, no switch is provided for it in the power supply. The +12 volts is turned off to most of the display (except the clock) and power to the relays is turned off for safety. The +5 volts remains on. Transformers T2 and T3 and capacitors C13 and C14 were removed from the Alliance rotator controllers.

mods to rotators

One small modification must be made to the rotators. Before removing the transformers and capacitors, connect the rotators to the Alliance control box and check to make sure they work properly. Turn the knob counter-clockwise until the rotator runs into the stops. Now open the rotator housing, but try not to disturb the gears. Tabs on the main shaft move a T-slug stop. Make sure the rotator has gone into the stop at full counter-clockwise rotation. Note the disc which activates the switch contacts; this produces the 10-degree signal. Carefully remove the snap ring, lift the gear, and position the disc as shown in **fig. 4**. Replace the snap ring and reassemble the body.

It probably would be a good idea to clean the contacts and lubricate the gears at this time. This disc is now at the 0-degree point and will produce a signal at every 5-degree set. I used a cable with 3 pairs of wires in aluminum foil wrap to connect the digital controller to the rotators. The two ground wires were used for AC return. Both rotator signals were brought down in one pair with the bare wire grounded at the controller.

A simplified version (manual operation, AZ-EL readout only) of the KA8OBL satellite tracker is available from the author. Send SASE to Rudolph Six, KA8OBL, 30725 Tennessee, Roseville, Michigan 48066

ham radio

a carrier-activated
CW reception limiter

Save your ears with this handy signal enhancer

Cut those big signals down to the same size as the small ones, the ad said, "use a limiter on your CW receiver." I did a double-take. Of course it's possible to do this — but only in the very special case where there's no more than one signal going through the limiter at any given time.

In the reality of our ham bands, this isn't always the case. CW bands usually present an irregular variation of anything from zero to several signals at any given moment in a typical audio passband. When two sine-wave signals are stuffed through a limiter, the stronger signal will suppress the weaker so that the difference between the two is nominally 6 dB greater at the limiter's output than at its input.[1] For example, if a pair of signals, one at 40 volts and the other at 2 volts, are fed to a 1-volt amplitude limiter, the 40-volt signal will be limited to 1 volt and the 2-volt signal will be reduced to 1/20 volt — 6dB, or 1/40th of a volt. Signals passed through an amplitude limiter are like Mother Nature's offspring — only the fittest survive.

Figure 1 shows what happens if, for example, random 4-volt and 2-volt CW signals are fed to a 1-volt limiter. For clarity, a filter that passes signal *B* and rejects the frequency of signal *A* is assumed. The unfiltered output of the limiter contains signals *A* and *B* plus harmonics and cross-products — a real mess.

Despite the poor quality of signal *B* at the limiter/filter output, experienced "brass pounders" can often get solid copy because of our fantastic ear-brain coordination capabilities. But comprehension is made more difficult, not easier, by the limiter's action. Moreover, if signal *A* is initially three or four times the amplitude of signal *B*, then copying *B* becomes very unlikely. Finally, even copying one signal can become a problem if heavy limiting is used. In this case, noise would come up to fill the spaces between code elements.

In addition to the suppression factor, if one or more signals above the limit threshold are fed to a limiter,

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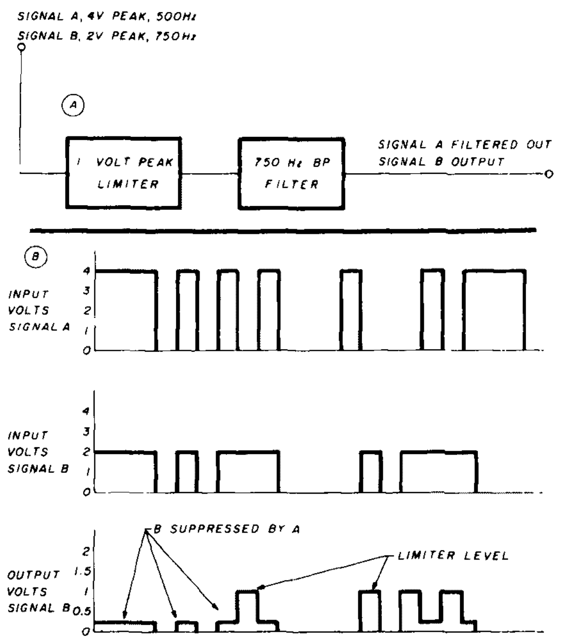

fig. 1. (*A*) limiter/filter circuit model; (*B*) impact of strong signal on weaker signal by limiter action. (Idealized peak envelope shown for clarity.)

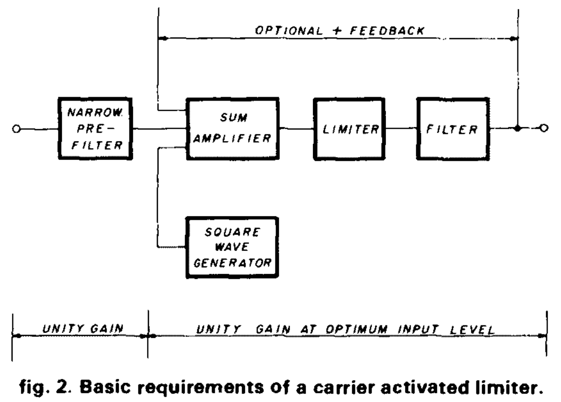

fig. 2. Basic requirements of a carrier activated limiter.

a squared-off waveshape, which contains a considerable amount of harmonic energy, is produced. In a recent article, a limiter was shown in front of a narrowband CW filter.[2] The article stated that locating the limiter in front prevents harmonics generated by the limiter action from reaching the power amplifier, prevents overload of the following power amplifier by strong signals, and makes up for AGC problems with CW. This is true as far as it goes, but it does nothing for the suppression problems illustrated above. In addition, it doesn't prevent sub-harmonics of the filter frequency from entering the limiter which, in turn, creates harmonics that appear as false signals in the filter passband.

To eliminate the problem with sub-harmonics and mitigate the suppression effect, a narrow filter must be added in front of the limiter. This practice prevents lower frequency signals that are sub-harmonically related to the output filter frequency from getting to the limiter and reduces the number of signals with which the limiter is likely to be confronted. But even with the addition of filters surrounding the amplitude *limiter, because of the isochronal* nature of CW, the* signal-to-noise ratio is worse when using a limiter than for the linear case. The S/N for the limiter case can, however, approach linear performance if a received signal can be adjusted to an amplitude level where limiting is just about to begin. (In this case, S/N refers to the ratio of the coded signals; ON response to the noise response during the coded signals' OFF state.) If the input level is made high enough so that receiver noise is at or above the limit level, then the effective S/N for any received CW signal will be 0 dB — usually an undesirable condition.

At this point one may wonder whether or not a limiter is worth the bother. Not only is the system no longer simple, but the input signal amplitude adjustment is very fussy, even if you're trying to get an S/N only nearly as good as would exist in the linear case — where the limiter could just be thrown out altogether. Fortunately something can be done — with just a little more hardware — that will tilt the S/N ratio more than 14 dB in favor of the limiter case.

the super-CW model 10: a carrier-activated limiter

Figure 2 shows a block diagram of what must be added to the basic Filter-Limiter-Filter combination in order to obtain the S/N enhancement. The basic concept recognizes that the noise floor, which is present between the coded elements, must be treated as a particular case. To accomplish control of this noise, a nominal 10-kHz squarewave carrier is linearly summed with the output from the pre-filter and fed to a low level limiter. The amplitude level of this carrier is made large enough to fully capture the basic limiter under no-signal conditions.

For a simplified explanation of operation, when a receiver's input to the system is set at a level where the noise floor output from the pre-filter is below the voltage amplitude of the carrier, the noise is suppressed *by* the carrier — remember, survival of the fittest. Then when a signal voltage appears that is above

**isochronal* — occurring at equal intervals of time

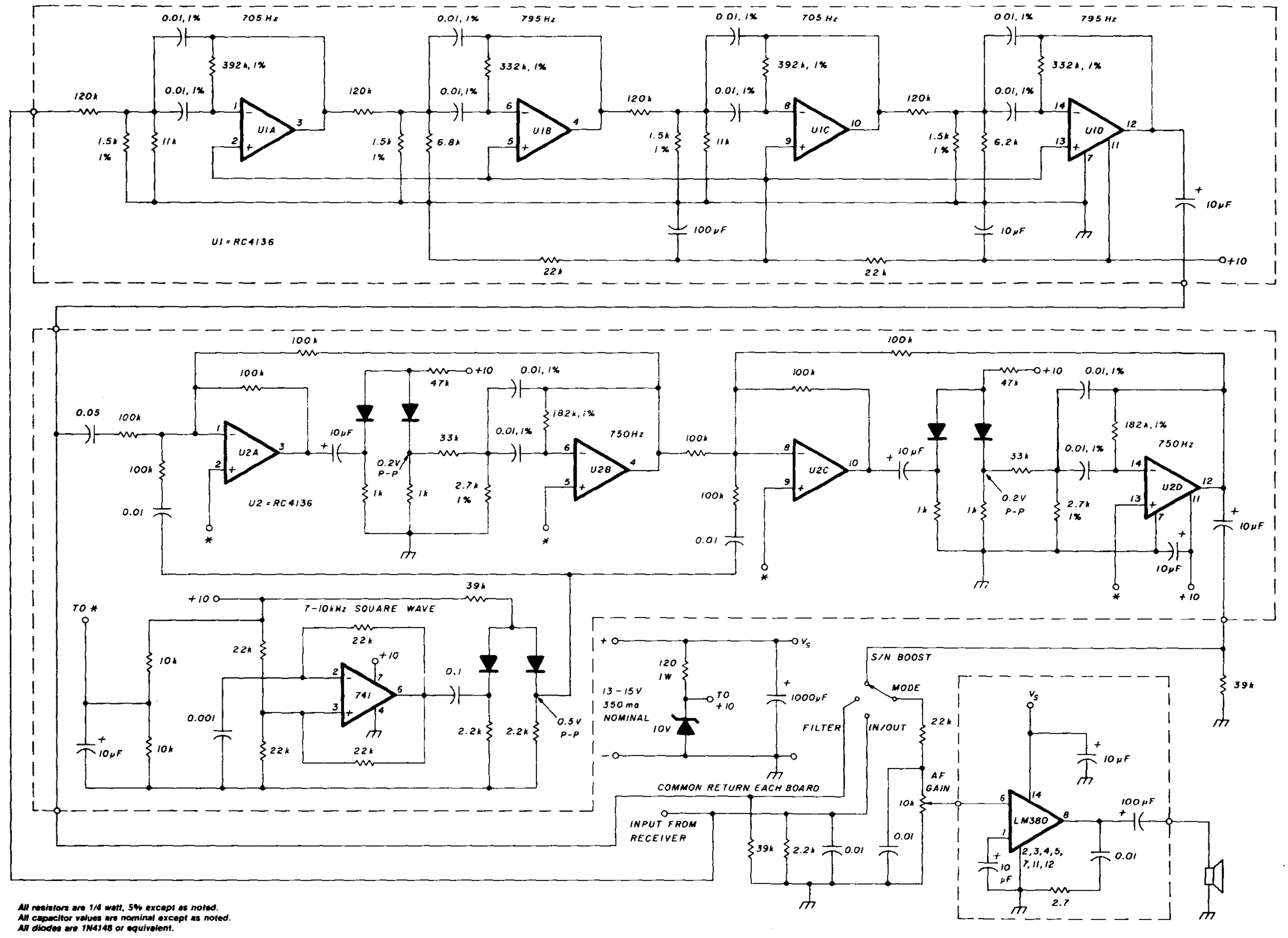

fig. 3. Complete system including pre-filter and a cascade of two carrier activated limiters with positive feedback and a power amplifier. (Patent applied for on part of this system.)

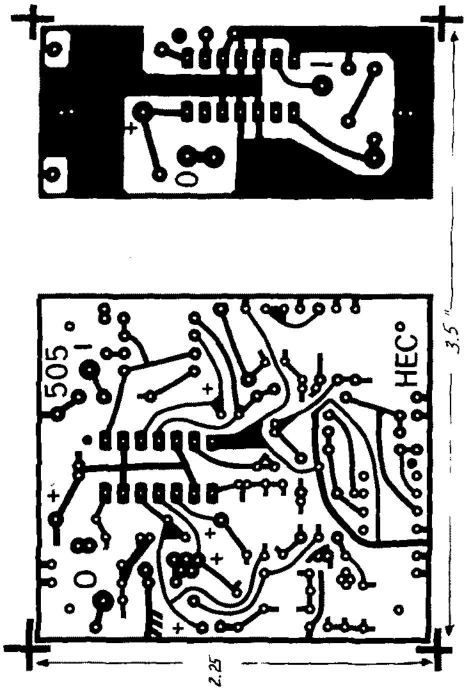

fig. 3A. PC boards 505 and 103, containing the carrier-activated limiter and the power amplifier (component side).

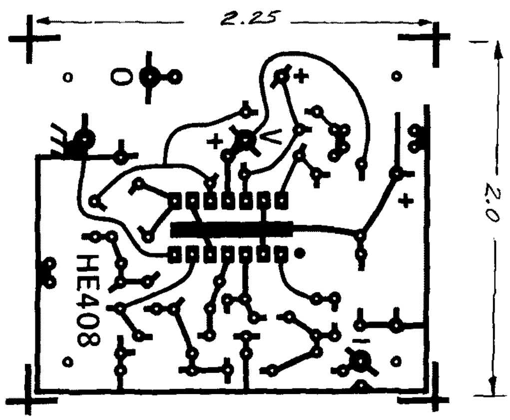

fig. 3B. PC board 408, the Butterworth filter board (component side).

the carrier amplitude, the signal suppresses the carrier. Moreover, the instantaneous *signal and carrier* voltages add to suppress coincident noise. The filter following the limiter removes harmonics created by the limiter as well as the 10-kHz squarewave carrier.

By designing the op-amp summer, limiter and op-amp active filter so that the gain is unity from input to output for the desired signal, identical stages may be cascaded for a multiplied S/N enhancement in a manner similar to that of a quality FM communication system. A complete system, including a narrow pre-filter and two stages of carrier activated limiters, which seems to provide just about the right amount of S/N enhancement for excellent operating "feel," is shown in **fig. 3**.

The input filter is *an 8th-order* Butterworth two-pole cascade with a 100-Hz bandwidth centered at 750 Hz. This design provides minimal ringing and much better skirt rejection than can be obtained from synchronously tuned filters or others of the Gaussian class. The summing amplifiers are designed to linearly add three inputs: the signal, the carrier, and a resistive controlled positive feedback, which assists the S/N enhancement of each stage. Circuit constants of the in-line limiters provide limiting levels of 0.2 volts peak-to-peak. The same limiter design is used to control output from the 10-kHz carrier, but in this case components are selected to make a 0.5 volt peak-to-peak limiter. Because of the light loading presented by the op-amp summer design, only one 10-kHz carrier generator/limiter combination is required to drive both carrier activated limiter stages. Identical 2nd order filters with a nominal 175-Hz bandwidth centered at 750 Hz are used following each carrier activated limiter. Output from the last filter is fed through an L network and volume control to drive an LM380 power amplifier. A mode control switch is included to enable selecting the input from a receiver, output from the linear pre-filter, or output from the complete carrier activated limiter system for input to the power amplifier.

system performance

With this system, S/N enhancement has essentially reached its maximum of a little over 14 dB when the S/N out from the pre-filter is +10 dB and enhancement declines smoothly toward zero as the S/N from the pre-filter approaches zero dB. As a relative comparison, this means that a received signal at about 4.5 dB *below* the noise in a typical 3 kHz bandwidth would yield a +10 dB S/N from the pre-filter and a +24 dB S/N from the carrier activated limiter system. **Figures 4A** and **B** illustrate generalized performance.

In addition to reducing the noise by 14 dB, the system can also suppress the ringing residues from the pre-filter by the same amount. Since the ringing residue amplitudes from a Butterworth filter are more than 10 dB below the ON elements of a CW signal that triggers them, the final effect is reduction of ringing residues to more than 24 dB below the desired signal.

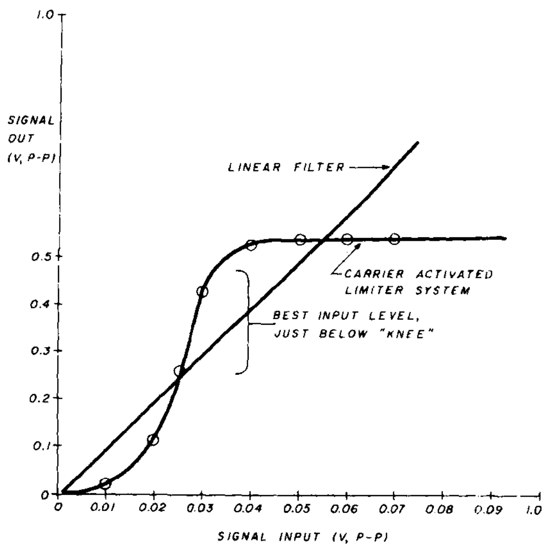

fig. 4A. S/N out versus S/N at input. Curve is analagous with S/N enhancement curves for FM systems, but better because this system has no detection losses.

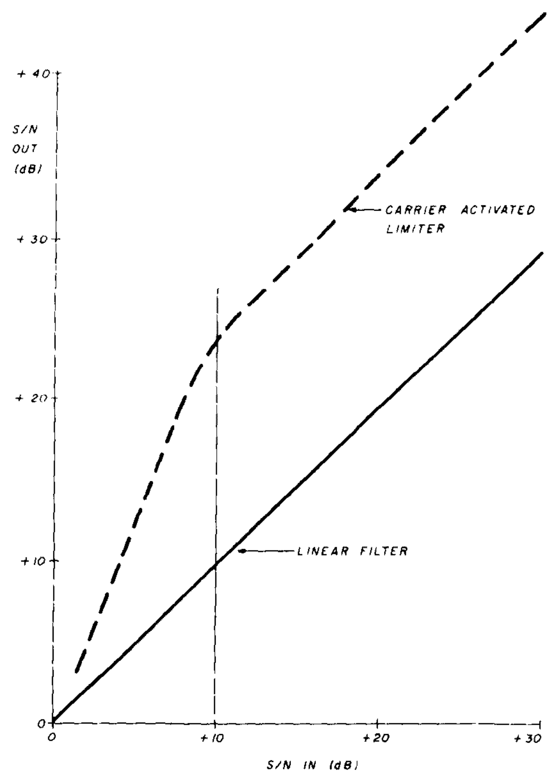

fig. 4B. Signal output characteristic as signal input level is varied. A nomianl 14 dB S/N ratio enhancement is available over the range from 0.2 to 0.5 volt p-p input with a peak boost of 18 dB when input is set at 0.3 volt peak-to-peak.

Also, because filter ringing increases in duration rather than amplitude as bandwidth decreases, this makes possible the use of sharper filters — down to 25 Hz or so for a CW signal at 20 WPM.[3] And this can be done without the need for a complicated synchronous detection system. However, to make use of this very narrow bandwidth and assure comfortable operation, many existing receivers would need better bandspread tuning control than they now have. Even the 100-Hz bandwidth used in this system requires relatively stable oscillators on the part of a received signal and the receiver as well as very careful tuning.

Since the limiter/filter system effectively holds the signal output level constant, the sound pressure listening level cannot exceed that selected by the AF gain control. This feature, present for any signal or noise pulse received, guards against damage to the listener's hearing. This protection is lost, however, when switching to the IN/OUT or Filter positions.

Other methods of S/N enhancement include those techniques that develop a voltage from a received signal to operate an electronic switch to turn an audio oscillator on and off, or to drive a digital system. In general, these methods are fine when a CW signal has a good S/N, but they usually don't perform well with threshold signals. However, the electronic switch method, where it must be used, can be improved by driving it from the carrier activated limiter system.

Finally, the system is designed to operate from a low input level to avoid the requirement of moderate to high receiver gain when listening to weak signals. This reduces the possibility of receiver saturation — unplanned limiting with all of its ramifications as shown in the beginning — by strong signals that may exist in a receiver outside of the pre-filter bandpass.

operating with the model 10

The nominal signal input voltage requirement is from 0.25 to 0.5 volt peak-to-peak at 700 to 800 Hz. Input termination is resistive at 2.2 kilohms, which enables the unit to be driven from either a receiver's speaker or phones output, or an op-amp. Output from the power amplifier is low impedance to enable driving a speaker of from 4 to 8 ohms. If headsets are used, connection to the output should be made through a nominal 100 ohms to avoid possible damage to ears or headset.

The MODE control has three postions. In the IN/OUT position, no filtering takes place, but the variable gain from the AF GAIN control is in effect. In the FILTER position, the 8th order Butterworth pair cascade is inserted between the input and the power amplifier. In the S/N BOOST position, the cascade of carrier operated limiters is connected between the FILTER output and the power amplifier.

To make the best marriage with a receiver, disable the AGC, use a low RF/IF gain, and set the audio level from moderately high to maximum. In a receiver that does not have an AGC ON/OFF capability, AGC action is effectively defeated by a reduced RF/IF gain setting. A little experience with the system will make this clear.

With the preliminaries completed, listen to the system with your receiver set at a point on its dial where no signal is present and with Model 10's AF GAIN set at mid scale. Then, switching through the three positions, you should hear broadband noise in the IN/OUT position, reduced noise in the FILTER position and still further reduced noise in the S/N BOOST position. As a quick reference, you can put Model 10 in its S/N BOOST position, then adjust your receiver's RF/IF gain control to obtain a weak noise background. Tune in a moderate-to-weak signal. Now, when adjusting your receiver's gain up and down, you'll notice a "knee" (see fig. **4B**) where the signal remains constant in level for increased input and where the signal falls off very sharply as gain is reduced. In the region of the knee, your receiver's gain control is adjusted for optimum S/N ratio enhancement. At this point, switching the MODE control to any of its three positions will result in about the same signal output level — which confirms an optimum receiver input level.

In the S/N BOOST position, tuning is even sharper than in the FILTER position. The BOOST position suppresses skirt responses and residual ringing in the same manner as it reduces the noise floor, which enables clean high speed copy with a narrow filter.

In normal operation you simply adjust receiver gain for input level and Model 10's AF GAIN for desired sound pressure level.

Don't want to build it yourself? The super-CW Model 10 is available preassembled. Contact the author at Hildreth Engineering, P. O. Box 60003 Sunnyvale, California 94088, for details.

references

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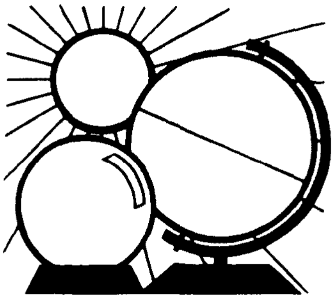
ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

equinoctial DX propagation brings change, excitement

This season's **propagation** is characterized by change and excitement resulting from the fewer number of daylight hours as winter approaches. In addition the MUF diurnal curve will no longer exhibit summer's typical low, flat response but will instead show a pronounced peak in the afternoon and a downward dip in the predawn hours. Short-skip sporadic E openings will become scarce, giving way to long skip, including many long one-hop transequatorial openings, particularly towards the south in the evening. Shortwave broadcasters must anticipate these changes by planning for frequency changes after each of the two-month periods, (March/April and September/October). The change is slower during the other two periods, which comprise four-month intervals (May through August and November through February).

During the equinoctial period the geomagnetic field exhibits the greatest variability. The magnetosphere (geomagnetic field) acts as a source of high latitude electrons and accounts for ionospheric movement. During equinox — when the earth's equatorial plane intersects the sun — solar wind particles feed down through the magnetosphere into both polar regions of the ionosphere. The variability, both in speed and density, of the solar wind stream is transferred to the geomagnetic field and ionosphere. Geomagnetic field control of the ionosphere is strong at this time; consequently we can look to it as a measure of what the ionosphere's reaction will be. Recall the large geomagnetic disturbances of last spring's equinox in April; the A index, which was 77 on April 21, was again up to 40 the following week, on April 28. Remember what propagation conditions were like then?

But what is meant by geomagnetic data? The A index is a measure of the amount of change, with time (minutes to hours), of the geomagnetic field from a normally (average) quiet day at that station. The larger the A index (one number per day) or K index (every three hours or 8 per day), the more the ionosphere is moving around. The eight K numbers are summed and multiplied by a factor that converts it into the daily single A index. (The number is further manipulated to calibrate the data from each measuring station.) Large changes in the ionospheric layers take place with K's above 6, especially for paths along or across the auroral zone (65 to 75 degrees north latitude). A K of 4 to 5 indicates moderate movement with its accompanying signal strength variation, QSB.

When listening on the bands during various degrees of geomagnetic disturbance, you'll notice something like this: with K's of 0 to 2 the band will be full of signals from just about every direction; with K's of 3 to 5, the accompanying absorption and some QSB will cut out some of the stations you hear. When the K's are 6 and 7, signals you hear will be coming from unusual, less frequently heard directions. For even higher K's, all signals are lost in the absorption and QSB. Don't despair — when the K drops slightly from the previous high, we're back in the "unusual direction" K category.

last-minute forecast

The third week of September is expected to be the best for working DX on the 10 through 30 meter HF bands, coincident with a higher solar flux. To confirm that this situation is occurring, monitor radio station WWV on 2.5, 5, 10, 15, and 20 MHz at 18 minutes after each hour. If the flux is up, expect transequatorial openings to occur to South Africa, South America, and especially Australia. These will be enhanced openings during a geomagnetic-ionospheric disturbance indicated by an A index of 20 to 40, which is likely at this equinoctial season of the year. The lower bands should greatly improve through the month with lower atmospheric noise levels and higher signal strengths. The MUF will be lower on east-west paths during disturbed conditions, with QSB noticed on the signals. Under these same circumstances, expect to hear DX from some unusual locations.

A full moon will occur on September 29th and its perigee on the 16th. The autumnal equinox will be on the 23rd at 0207 UTC. No significant meteor showers are expected this month to enhance meteor burst DXing.

band-by-band summary

Six meters may have a few sporadic E openings around local noon, but don't count on them. This month should be the last chance for E_s until next summer's season begins.

Ten, twelve, and fifteen meters should provide a few short-skip E_s openings and many long-skip openings during any solar flux peaks to most southern areas of the world during daylight. Some of these openings will result from transequatorial propagation, especially during the disturbed conditions.

SEPTEMBER

WESTERN USA

GMT	2300	2200	2100	2000	1900	1800	1700	1600	1500	1400	1300	1200	1100	1000	0900	0800	0700	0600	0500	0400	0300	0200	0100	0000	
PDT	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	
ASIA FAR EAST	20	20	40	40	40	40	40	30	30	30	30	40	40	40	40	40	30	30	30	20	20	20	20	20	N
EUROPE	30	30	20	20	20	20	20	20	20	20	20	20	30	40	40	40	40	40	30	40	40	40	40	40	NE
S. AFRICA	20	15	15	12	10	10	10	10	10	10	12	15	20	20	20	20	20	20	20	20	20	20	20	20	E
S. AMERICA	10	10	10	10	10	10	10	12	12	15	20	20	20	20	20	20	20	20	20	20	15	15	12	10	SE
ANTARCTICA	12	12	12	12	15	15	15	20	20	30	30	30	30	30	30	30	30	20	20	15	15	15	15	12	S
NEW ZEALAND	10	10	12	12	15	15	20	20	20	20	20	20	20	20	20	15	15	12	12	10	10	10	10	10	SW
OCEANIA AUSTRALIA	10	10	12	15	15	20	20	20	20	20	20	20	20	20	15	15	15	12	10	10	10	10	10	10	W
JAPAN	15	15	15	20	20	30	40	40	30	30	40	40	40	30	30	30	30	30	20	20	20	15	15	15	NW

MID USA

MDT	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	
ASIA FAR EAST	30	30	40	40	40	40	40	30	30	20	20	20	20	20	20	40	40	40	40	40	30	30	30	20	N
EUROPE	30	30	20	20	20	20	20	20	20	20	20	20	20	30	40	40	40	40	40	40	40	40	40	40	NE
S. AFRICA	20	15	15	12	10	10	10	10	10	10	10	12	15	20	20	20	20	20	20	20	20	20	20	20	E
S. AMERICA	10	10	10	10	10	10	10	10	12	12	15	20	20	20	20	20	20	20	20	20	15	15	12	10	SE
ANTARCTICA	15	15	12	12	15	15	15	15	20	20	30	30	30	30	30	30	30	30	30	20	15	15	15	15	S
NEW ZEALAND	10	10	12	12	15	15	20	20	20	20	20	20	20	20	20	20	20	20	15	12	12	10	10	10	SW
OCEANIA AUSTRALIA	10	10	12	12	15	20	20	20	20	20	20	20	20	20	20	20	20	20	20	15	12	10	10	10	W
JAPAN	20	15	20	20	20	30	40	40	40	40	30	30	30	30	40	40	40	30	30	30	30	20	20	20	NW
CDT	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	

EASTERN USA

EDT	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	
ASIA FAR EAST	30	30	40	40	40	40	40*	40	30	20	20	20	20	20	20	40	40	40	40	40	30	30	30	20	N
EUROPE	40*	30	30	20	20	20	20	20	20	20	20	20	20	20	20	30	40	40	40	40	40	40	40	40	NE
S. AFRICA	20	15	15	12	10	10	10	10	10	10	10	10	12	15	20	20	20	20	20	20	20	20	20	20	E
CARIBBEAN S. AMERICA	10	10	10	10	10	10	10	10	10	12	12	15	20	20	20	20	20	20	20	20	15	15	12	10	SE
ANTARCTICA	15	15	15*	15*	15	15	15	15	15	20	20	30	30	30	30	30	30	30	30	20	15	15	15	15	S
NEW ZEALAND	10	10	12	12	15	15	20	20	20	20	20	20	20	20	20	20	20	20	20	15	12	12	10	10	SW
OCEANIA AUSTRALIA	10	10	12	12	15	20	20	20	20	20	20	20	20	20	20	20	20	20	20	15	15*	12	10	10	W
JAPAN	20	15	20	20	20	30	40	40	40	40	30	30	30	30	40	40	40	40	40	30	30	30	20	20	NW

The italicized numbers signify the bands to try during the *transition and early morning hours*, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

Twenty, thirty, and forty meters will support propagation from most areas of the world during the daytime and into the evening hours almost every day, either long-skip to 2500 miles (4000 km) or some short-skip E_s to 1250 miles (2000 km) per hop.

Thirty, forty, eighty, and one-sixty meters are all good for nighttime DX. The bands will be open in the east soon after sundown, swing toward the south about midnight, and end in the Pacific areas during the hour or so before dawn. On some nights these bands will be as good as the winter DX season. Some nights, noise (QRN) from thunderstorms (airmass and fronts of fall weather storms) will be a problem. However, high signal levels that occur during short-skip E_s propagation conditions may overcome the static.

ham radio

the
weekender

a simple continuity tester

A United States Navy training manual once described a circuit — affectionately known as the "octopus" — for examining semiconductor junctions, capacitors, and comparing printed circuit boards similar to the circuit shown in **fig. 1**. Through the years, I have run across several other versions of the octopus differing only in the methods of attaching the many wires connecting the instrument to the oscilloscope, power source, and device under test. The octopus is useful in pointing out problem nodes and providing fast go/no-go tests on a variety of components. Oscilloscope patterns obtainable are combinations of those depicted in **fig. 2**.

While ferreting out the cables and cords necessary to hook up a particularly elaborate octopus, it occurred to me that I would not be wasting time looking for a power cord if the octopus were battery powered. After locating the power cable, I determined that elimination of the required oscilloscope would also be necessary because every available oscilloscope was hopelessly buried under other people's ongoing calibration setups.

circuit description

As a result of my decision to eliminate both the power cord and the oscilloscope, the circuit shown in **fig. 3** was developed. The circuit illustrates a 555 timer operating as a minimum-parts astable oscillator. The ceramic sounder provides an aural indication of resistance. The diode bridge steers current through the ceramic sounder with the correct polarity, and the dual color LED (light emitting diode) provides a visual indication of current flow and polarity information. The tapped power source provides a node that is negative when the 555 timer is conducting and positive when the timer is not conducting. Current through the LED and the device under test is limited to a nondestructive value by the internal resistance of the ceramic sounder.

By Robert R. Frahm, WD6GMB, Box E, American Embassy, APO Miami 34002

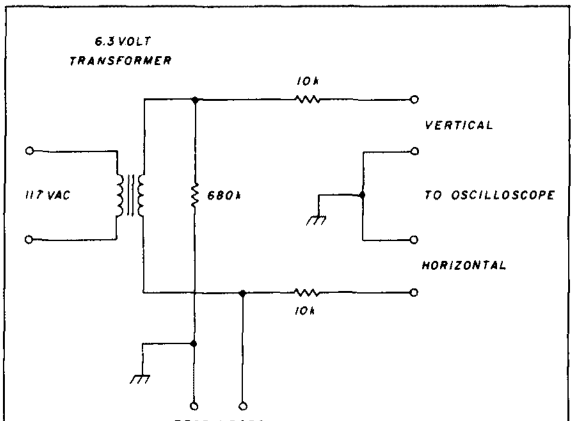


fig. 1. The "octopus."

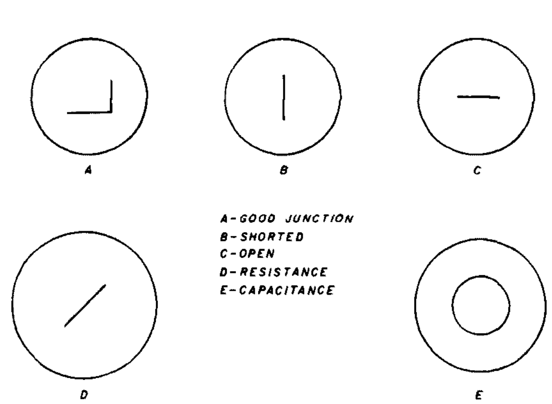


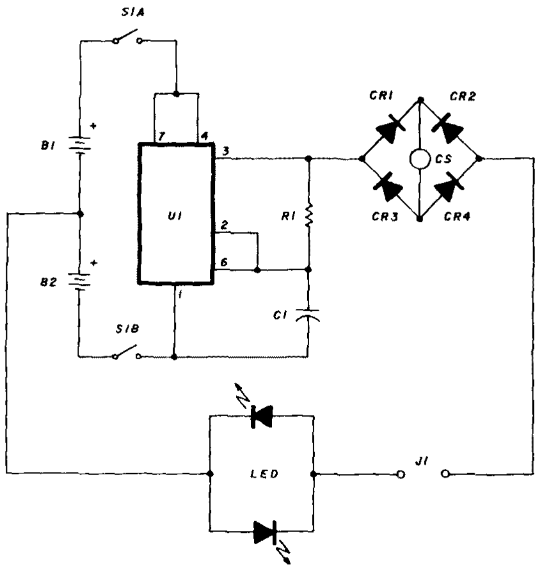
fig. 2. Octopus presentations.

directions for use

With the junction tester turned on, a *shorted* condition is indicated by a steady tone and by the alternately blinking red and green LED. This condition can be observed by holding the test probes together. The tone should be steady — i.e., there should be no change in volume.

An *open* condition is indicated by silence and the absence of color in the LED. A clicking noise may be heard if the test probes are held close together due to capacitance between the probes.

A good junction is indicated by a beeping sound and either a blinking red or blinking green LED when the test probes are placed across the junction. If the test probes are reversed, the same sound will be heard, but the LED will blink in the opposite color! For ex-



B1,B2	9-volt battery, RS No. 275-1546
C1	2.2 μF 50-volt capacitor RS No. 272-997
CR1,2,3,4	1N914 diode, RS No. 276-1620
CS	ceramic sounder, RS No. 273-060 or No. 273-065
J1	female banana jacks, RS No. 274-725
LED	XV-5491 LED (dual color), RS No. 276-035
P1,P2	battery connectors, RS No. 270-325
R1	33 kilohm, 1/4-watt resistor, RS No. 271-1341
S1	DPDT switch, RS No. 275-1546
miscellaneous	box, RS No. 270-222
	alligator clip, RS No. 270-347
	test probes, RS No. 278-704
	printed circuit board, RS No. 276-168
	IC socket, RS No. 276-1995

Note: RS = Radio Shack

fig. 3. Junction tester.

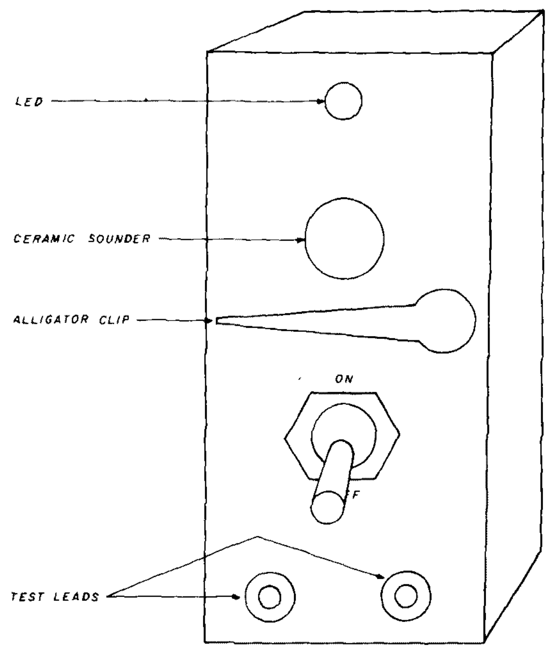


fig. 4. Panel layout.

ample, if a beeping sound and blinking red LED indicate a good junction, then a beeping sound and blinking green LED indicate that the test probes are reversed. Take note of this relationship, since it helps to determine the polarity of the junction.

The junction tester may be used both in and out of the circuit (with the power off), but parallel resistances may complicate interpretation of the results. Comparing the results of testing printed circuit boards known to be good can facilitate interpretation of the results of testing boards of unknown status.

Low current pulses (10 Hz at 18 volts peak-to-peak) are available at the test probe tips. These signals are useful when implementing signal or pulse injection techniques.

A slight change in volume while the test probes are shorted may indicate that the batteries are "out-of-balance." This condition may be corrected by interchanging or replacing the batteries. Since the junction tester draws current from the batteries whenever it is turned on, be sure to turn it off after use.

construction details

Circuit layout is not critical. A socket is recommended for the 555 timer to simplify replacement. The low current (CMOS) 555 timer version is unforgiving of inductive loads and not recommended, particularly on modified junction testers that may include a switch to bypass the ceramic sounder and LED. A suggested panel layout is presented in **fig. 4**. The total cost of parts is about $20.00 if they're purchased new.

conclusion

An alligator clip secured to the panel holds a loose transistor, sparing me the bother of chasing the transistor around the bench with the test probes. (Since building the junction tester, I still find myself searching for it because other technicians are always carrying it away.)

I've often found faulty parts while demonstrating the junction tester. Once I used it to "wring out" a problem extender board that had been tested with an ohmmeter, but showed no apparent defects. Using the tester, I found a staked terminal where oxidation had developed between it and the printed circuit, turning the mechanical connection into a diode!

ham radio

product REVIEWS

world time clock

Azimuth Clock has just released its new World Time/Dual Zone clock. Using a large 1-7/16 × 2-1/2-inch LCD display, it's easy to read. And with large pushbuttons on the side, it's also easy to set. There's no need to fumble with a pen or pencil on tiny buttons recessed into the clock's mechanism.

To ensure accuracy, these clocks use a quartz crystal time base oscillator. The local time display shows time in the standard AM/PM format. The Universal Time clock display is in 24 hour/Zulu/military notation for instant logging accuracy. Both displays can be set to also flash the date, if desired. Both the UTC and local time clocks have been programmed to automatically determine the number of days per month and will require resetting only during leap years, or should the battery fail.

These clocks are set in a durable piece of extruded aluminum that has been beautifully silk screened. Azimuth also offers a 14-day satisfaction-guaranteed or your money-back offer: if you're not delighted with your Azimuth clock, simply return it within 14 days of purchase. The World Time dual zone clock retails for $29.95 but is currently specially priced at $24.95, plus $1.95 shipping and handling.

Toll-free numbers are available for ordering by credit card; call (800) 821-6842; in California, (800) 421-1061. Further information is available from Azimuth, 11030 Santa Monica Blvd., Suite 200, Los Angeles, California 90025.

Circle #307 Reader Service Card.

N1ACH

CMC's docking booster

Here's a really neat product for owners of handheld radios. The Docking Booster is a compact, self-contained power supply, a 16-dB gain GaAs MES FET receiver pre-amp, and a 30-watt RF amplifier (a 50-watt unit will be available soon), all in one package. Designed to integrate your portable radio into your car without sacrificing either portability or versatility, the Docking Booster is currently available for the IC-2AT, IC-02AT and Yaesu, Kenwood, and Standard radios.

When I unpacked the box, I was surprised that the unit is really quite small. Installation is quite simple. You just connect the unit to the car battery either directly or through the fuse block, slide the radio in, attach the external antenna and you're on the air!

The Docking Booster comes with a U-shaped bracket that can be attached to the door of most cars. (Unfortunately, our Dodge station wagon door is too deep, and the unit wouldn't fit there. I also found that both the power cord and antenna feedline tend to get in the way if you install the radio on the driver's door.) But I did find the perfect place to install the unit — right in the ashtray (I don't smoke). The U-clamp which fit perfectly, provides a stable, easy-to-reach mounting platform.

Now, trying to shout into the ashtray while you're driving down the road is bound to attract attention, if not cause an accident. So you'll want to add the optional ICOM handheld speaker mike.

The Docking Booster is simple to use, with only a high and low power switch for the 30-watt RF amplifier. There are no other switches, knobs, or dials to turn or set.

Here in Southern New Hampshire, I found the extra 30 watts to be helpful in accessing repeaters. The 16-dB pre-amp helped tremendously in hearing distant repeaters and weak mobile stations on simplex frequencies.

I did find one problem with the unit. You have to avoid long-winded transmissions because there's no heatsink for the RF amplifier. I don't yet have enough experience with the unit to tell whether this is going to be a problem or not, but in normal use, the Docking Booster did get warm to the touch. (The 50-watt unit does have a heat sink.)

The Docking Booster uses the latest Japanese power transistors in a stripline circuit for high, stable power output. Using a 1.5-dB NF GaAs FET device, the 16-dB receiver provides excellent weak signal reception over the full 2-meter band. Designed to run on 13.8 volts DC, the unit requires 4 amps maximum on transmit and supplies a stable 10 volt power source for the radio.

For more information on the Docking Booster, contact CMC Communications, 5479 Jetport, Tampa, Florida 33614.

N1ACH

Circle #305 on Reader Service Card.

NEW products

integrated packet terminal

Packeterm, a company specializing in packet radio technology, has announced release of the PACKETERM IPT. The IPT is a complete packet node controller and terminal in one compact unit. The IPT will interface with any popular FM or SSB rig, for use on both HF and VHF channels. The IPT can operate at 300 or 1200 baud. All that's required for operation is to connect the IPT to a transceiver and tune to one of the local packet frequencies.

The IPT uses Tuscon Amateur Packet Radio

firmware for complete compatibility with existing packet networks. A built-in modem allows much greater tolerance for receiver audio level than previous available designs.

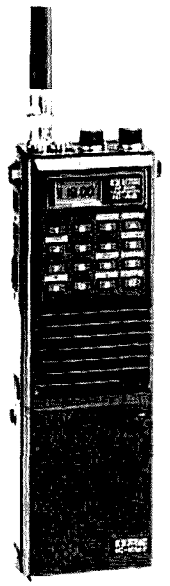
The IPT will also function as a general-purpose ASCII terminal. It features data rates from 110 to 19,200 baud, 7 function keys, 40 or 80 column display, and will drive an optional printer. Power required is 120 VAC at 1/2 ampere. For portable operation, a DC converter that uses a 13.8 VDC power source is available.

For further information, contact Packeterm, P.O. Box 835, Amherst, New Hampshire 03031.

Circle #303 on Reader Service Card.

ICOM IC-A2 air band handheld transceiver

ICOM has announced the IC-A2 5-watt PEP output aircraft handheld transceiver. Utilizing over 20 years of experience in synthesized communications equipment, the ICOM IC-A2 in-



cludes outstanding new features not available in other handhelds.

Standard features include all 720 COM and 200 NAV channels plus 720 additional COM channels and 200 NAV channels. Five watts PEP power *output — an ICOM exclusive —* and 1.5 watts operation capability for battery-saving low power operation are standard.

The unit features ten owner-programmable *memory* channels, with internal lithium cell memory backup.

The ICOM IC-A2 comes standard with an IC-CM7 rechargeable Nicad battery pack, charger, LC-14 soft leather case, and earphone. A wide selection of options and accessories are available, including the ICOM HS-10 headset and HS-10SA VOX unit or HS-10SB PTT switchbox.

The ICOM IC-A2 may be purchased only through authorized ICOM avionics dealers/distributors or FBO shops.

For further information, contact ICOM, 2380 116th Avenue N.E., Bellevue, Washington 98004.

Circle #306 on Reader Service Card.

packet radio relay power amplifier

With linking of packet radio stations on 220 MHz becoming popular, Hamtronics has announced a new version of their widely-used power amplifiers designed just for this service. Called the PPA-220, the new PA is similar to the Hamtronics LPA 2-40, except that it has increased gain (up to 50 watts out with 2 watts drive) and built-in PIN diode antenna switching to allow T/R transition in only a few milliseconds. The new PA can be used with the Hamtronics T51 or any other 2 watt, 220 MHz exciter. With the ultra-fast T/R switching, the PPA-220 is an ideal component for packet relay stations being constructed for inter-area ties at 9600 baud. The cost of the kit is $138.00.

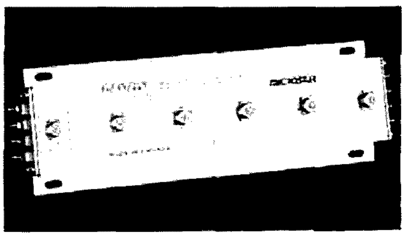
For more information on this and other products, such as transmitters, receivers, repeaters, and "202" type modulators and demodulators, contact Hamtronics, Inc., 65 Moul Road, Hilton, New York 14468-9535.

multi-tap V/H switch

The MTS-1200 is the first in Gensat's series of MICROSTAR® integrated block downconversion accessories. Incorporating all the circuitry required to independently operate up to four receivers from one dish, the MTS-1200 offers the easiest solution to most MTVRO system requirements.

The MTS-1200 is designed to operate with most block downconversion receivers operating at 950 to 1450 MHz. Each receiver can independently select the vertical or horizontal line from the block downconverters. No signal loss is introduced by the accessory.

The switching circuitry is driven from the polarizer output of the receiver. In order to ensure proper isolation between the horizontal and



vertical signal, the polarizer interface must be compatible with either a ferrite or pin diode polarizer.

Packaged in Gensat's standard weather-proof housing, this MICROSTAR® product can be mounted either indoors or outdoors.

For further information, contact Gensat, 951 Alness Street, Downsview, Ontario, Canada M3J 2J1.

Circle #308 on Reader Service Card.

NEW products

digital level meter

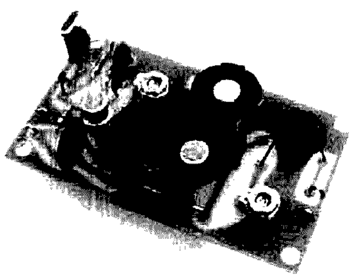
North American SOAR Corporation has announced the release of its Model 1700 digital level meter, which represents an entirely new concept in surface level measurements. Model 1700 was designed to assist the professional engineer or craftsman in making finite level or angular measurements quickly and accurately. Small and compact, the unit is sized 6-1/4 × 3-3/4 × 1-1/4 inches for portability. The Model 1700 has two 3-1/2 digit LCD readouts with annunciators, enabling the operator to observe the slope angle in degrees, the slope direction, and slop height in meters per meter from either the side or top. A buzzer can be activated to give an audible tone alarm when 0 degree (level) is attained. It contains a switchable LCD light for reading in dim light. Powered by two 1.5 volt "AA" batteries, the Model 1700 can be continuously operated for 500-800 hours. The measuring method employs an encoder and micro-computer, giving it a dynamic range of ± 120 degrees with an accuracy of 1.0 degree and resolution to 0.5 degree. The Model 1700 is priced at $99.95 in small quantities.

For more information, contact North American SOAR Corporation, 1126 Cornell Avenue, Cherry Hill, New Jersey 08002.

Circle #310 on Reader Service Card.

transmitter kit

RF Kit Co. has developed a compact Amateur Radio transmitter kit that can be assembled in one evening and provide a solid 5-watt CW output on the 7, 10.1, and 14 MHz Amateur bands. This unit is crystal controlled and one crystal is included in the $39.95 price. Stock crystals available include: 7.035, 7.125, 10.108, 10.125, and 14.035 MHz.



This unit is suitable for home use with a simple dipole antenna or for portable and emergency power communications.

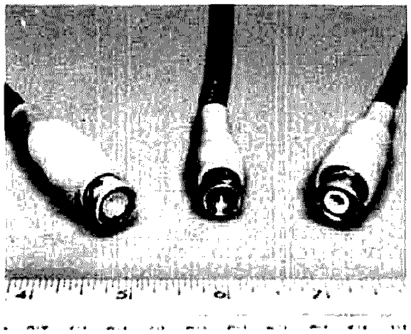
The electronic design is straightforward and may be easily modified for any Amateur band from 160 to 10 meters. The 3.5 × 2 × 1 inch size will allow the builder to custom-fit this transmitter into many surplus enclosures.

For further information, contact RF Kit Company, P.O. Box 27127, Seattle, Washington 98125.

Circle #302 on Reader Service Card.

weather boots

Kilo-Tec has announced a new custom weather boot for use with RG-58, RG-59, and RG-8X. Simply slip the weather boot over the



coaxial cable before soldering on the connector, then slide the boot over the connector for a good weather-tight seal — without the use of tape or rubber compounds.

The boots are manufactured with a flexible vinyl material that resists moisture and breakdown from the sun's rays.

Three new boots are available: Model KTBNC-59 for (F) BNC/RG-59 and RG-8X; Model KTBNC-58 for (F) BNC/RG-58; and Model KTBF-59 for (M) Type F/RG-59 and RG-8X. We also offer various models for PL-259 and type N. Boots for TNC are available on special order.

For more information, contact Kilo-Tec, P.O. Box 1001, Oak View, California 93022.

Circle #313 on Reader Service Card.

RF circuit design program

STAR 1.0 is a low-cost program written for circuit design using personal computers. It analyzes and optimizes electronic circuits including amplifiers, oscillators, filters, matching networks, hybrids, couplers and others. STAR 1.0 features include frequency domain analysis of circuits, optimization of any component values in circuit, file storage of circuits for easy recall or modification, and screen, printer, and plot outputs. The program disk includes 30 application examples. The program is available for IBM. PC/XT/JR, APPLE II+IIc/IIe, Kaypro 2/2X/4/10 CP/M and the Commodore C-64.

For additional information, write Circuit Busters, P. O. Box 256, Lilburn, Georgia 30247.

Also available from Ham Radio's Bookstore — *IBM and C64, $99; Apple and Kaypro, $89.* Add $3.50 for shipping and handling.

Circle #309 on Reader Service Card.

new antennas

NCG Company has announced its new line of Amateur antennas. The line includes a 1.2 GHz antenna for base/repeater use (with a 12.5 dB gain) and a new mobile with a 7.5 dB gain.

Also available are the new 2-meter 70-cm antenna duplexer and 6-meter/2-meter antenna duplexer combinations. A quad band vertical for HF + VHF 40, 15, 10, and 6 meters is ideal for the limited roof areas.

Another new antenna is the new ground-plane vertical Tribander for the new 10, 12, and 18-MHz band.

For further information, contact NCG Company, 1275 North Grove Street, Anaheim, California 92806.

Circle #312 on Reader Service Card.

CoCo MORSE

The latest release from dataLOG Software allows you to send morse code from your Radio Shack Color Computer. "CoCo MORSE" requires 32K of RAM and a disk drive. It features a 320-character input buffer for type ahead that will keep up with the fastest typist. (The operator is usually not aware of the buffer limit because the buffer works in a circular fashion, allowing wrap-around when the 320 character limit is reached.) A split screen display shows input buffer and transmitted characters.

The program will allow operation from 5 to 80 words per minute and speed changes may be made during transmission. Up to five predefined message buffers can be created and saved to disk for increaed efficiency during contests or for that "brag" tape. The messages may be dumped into the transmit buffer, up to the buffer's limit, without effecting transmission speed or quality.

Included in the purchase price is an RS232 interface to connect the computer to the transmitter. The interface also provides the necessary isolation between the transmitter and computer. CoCo MORSE is written in machine language and is relocatable. Its documentation includes information for modifying the weight and duration of the transmitted code. A future release of the "dataLOG Logbook" program will allow combined operation of the morse program AND the log program for the dedicated CW operator. CoCo MORSE, including the interface sells for $39.00 plus $2.50 for shipping and handling.

For additional information, contact dataLog Software, P.O. Box 10531, Jacksonville, Florida 32247.

Circle #311 on Reader Service Card.

antique radios

Antique Radio Classified, a national publication for buyers and sellers of old radios and related items, is entering its second year of publication. The largest publication of its type, it caters to collectors interested in collecting radios

built between 1905 and the 1940's. Although the emphasis is on "Wanted" and "For Sale" ads, many articles related to the hobby are included in each issue. Every subscriber is entitled to a free classified ad in each issue.

Antique Radio Classified is published 12 times a year in an attractive 5-1/2 × 8 inch size format.

For further information, contact Gary B. Schneider, publisher, Antique Radio Classified, 9951 Sunrise Blvd., Suite R-9, Cleveland, Ohio 44133. (Enclose SASE.)

Circle #315 on Reader Service Card.

'tenna hitch

Mounting mobile antennas on cars has always been a problem. Most people are reluctant to drill holes into the car body, and now that many bumpers are made of plastic, that relatively easy way out is no longer usable, either. However, DC Sales has come up with an idea that makes mounting antennas on cars equipped with trailer hitches as easy as 1, 2, 3!

The 'Tenna Hitch is designed to fit underneath the trailer hitch ball and is equipped with the standard 3/8-inch insulated mount. The 'Tenna Hitch is made from high strength, chrome plated steel and is designed to give years of trouble-free service. It will mount all currently available commercial mobile antennas.

For more information, contact DC Sales, 1602 Chestnut Ridge Road, Kingwood, Texas 77339.

Circle #314 on Reader Service Card.

2-meter RF amplifier

Hustler has announced the availability of its new model HVA-225 Class C Amplifier for 144-149 MHz 2-meter FM Amateur mobile use. Utilizing state-of-the-art broadband microstrip design, the amplifier exhibits components and manufacturing techniques found only in higher priced commercial units for superior performance and reliability.

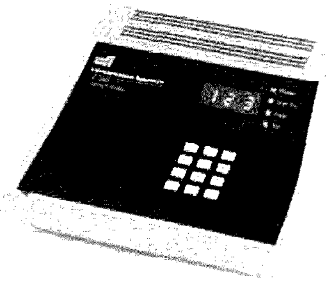
The HVA-225 is conservatively rated at 25 watts with only 2 watts of drive while requiring only 4 amperes at 13.8 VDC for full output. Separate power and RF indicators, on-off switch, SO-239 connectors, reverse polarity protection plus extra capacity heat sink for high temperature reliability and efficiency. The amplifier is housed in a black matte finish aluminum housing complete with gimble bracket and thumb screws for underdash mounting.

For more information on the HVA-225 amplifier, write Hustler, Inc., 3275 North "B" Avenue, Kissimmee, Florida 32758.

Circle #317 on Reader Service Card.

1000-call paging encoder

Communications Specialists has just announced the availability of its new PE-1000 desk-top paging encoder. The PE-1000 is capable of 100 or 1000 call paging capacity for two-tone sequential signaling. Five-tone sequential and REACH formats are also available. All features are included in every unit and fully field programmable through the front panel keyboard. A non-volatile memory retains the programming if a power loss occurs.



The PE-1000 includes all standard tone groups from 250.0 Hz to 4000.0 Hz and uses state-of-the-art microprocessor technology for high accuracy and stability of paging tones. The unit provides outputs for recording a hard copy print-out of all paging activity and an automatic self test is run each time the PE-1000 is powered up. The PE-1000 is protected by a full one-year warranty and is available for immediate shipment from factory stock. The price if $224.95.

For information, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665.

Circle #316 on Reader Service Card.

communications satellites

by Larry Van Horn

The first comprehensive directory of communication satellites and their radio frequencies is now available from Grove Enterprises.

Chapters cover spy and surveillance satellites, U.S. and Russian manned space missions, military tactical and scientific satellites, oceanographic and water satellites, navigational and communication satellites, and private and direct broadcast satellites.

This directory of space communications includes chapters on channelization band plans, transponder identification, international satellites, and a history of earth satellite development. An exhaustive frequency cross-reference allows quick identification of the source of unknown transmissions from spacel

Illustrations and tables are included for better understanding of this space technology. Special chapters provide insight into satellite operation, much of which has never been revealed to the public.

The price is $12.95. Book-rate shipping by mail is free. (For shipping by UPS, add $1.50.)

To order, contact Grove Enterprises, P.O. Box 98, Brasstown, North Carolina 28902.

Also available through Ham Radio's Bookstore, $12.95 plus $3.50 shipping and handling.

Circle #301 on Reader Service Card.

NEW products

DeVry to coordinate nationwide VE program

The DeVry Technical Institute of Technology s planning to expand its VEC coverage from the 9th call district to all 13 districts according to Jim Georgios, W9JUG, DeVry VE Coordinator. DeVry is looking for groups of three Extra Class Amateurs to set up "core groups" that will function as DeVry representatives in their local areas. DeVry will provide core group members with three written exams per license class and a computer generated tape for code tests. The core groups will then be responsible for setting up, administering, and correcting of exams. DeVry requires that core groups comply with FCC Part 97, Subpart I rules and regulations.

Initially DeVry representatives will schedule exams at least every other month, though a once-monthly schedule is preferred.

If your club or group is interested in working with the DeVry Institute of Technology, call Jim Georgios at (312) 929-8500 between 9:00 A.M. and 4:00 P.M. (CDST) or write to him at DeVry Institute of Technology, 3300 N. Campbell Avenue, Chicago, Illinois 60618.

Circle #318 on Reader Service Card.

advanced terminal unit

Said to be the most advanced HF RTTY modulator-demodulator available, the ATU-1000 from AEA is designed for interconnecting a communications transceiver with a computer (or mechanical communications device) using appropriate computer communications software. Operation on Morse, Baudot, ASCII, Packet and AMTOR teletype is provided.

The heart of the demodulator system of the ATU-1000 is a pair of identical eight pole (0.5 dB ripple) Chebyshev filters. Twin oscillator-modulators mix the input tones to the filter's center frequency.

Both mark and space tones may be adjusted independently from 1000 to 3000 Hz, providing compatibility with all commercial and Amateur tone pairs. Adjustment of the tone frequencies is accomplished with ten turn potentiometers and the tone frequencies are indicated with a built in frequency counter. Adjustment of the filters is accurate to 1 Hz. For fixed tone pair operation, an optional eight-pole bandpass prefilter is selectable from the front panel. Selection of the normal channel bandwidth of 180 Hz or a narrow 100 Hz bandwidth is provided for optimizing the channel filters for the mode of operation.

The filter system is followed by twin full-wave detectors and twin four-pole low-pass filters with selectable cutoff frequencies corresponding to 50, 110, and 300 baud data rates. This is followed by a DC coupled threshold correction system that provides superior performance during selective fades and low signal conditions. In total, 32 poles of receive filter are provided in this system.

Operation using both normal mark-space comparison and mark only or space only is selectable during interference on one of the information channels.

Tuning is indicated with a discriminator style LED bar graph with selectable mark only, space only and summed mark and space operation. In addition, the tuning rate is selectable for quick initial tuning and precise final tuning.

For further information, contact Advanced Electronic Applications Inc., P.O. Box C-2160, Lynnwood, Washington 98036.

Circle #304 on Reader Service Card.

Heath TNC

The state-of-the-art HD-4040 Terminal Node Controller (TNC) from Heath is said to be the only RFI approved TNC on the market.

A version of the popular Tucson Amateur Packet Radio (TAPR) TNC, the Heath TNC allows communication using terminal or computer control of any Amateur Radio system. Packet radio insures error-free communication and greatly increases communication speed. The HD-4040 has a built-in 1200 baud modem, although Baud rates up to 9600 are possible with an external modem. Both AX.25 and VADCG protocols are used.

Three modes of operation are provided: a conversation mode that allows conversation with another operator, a command mode that allows configuration of the TNC and use of variety of operating commands, and a transparent mode that's used in transferring files from one computer to another. A 6809 processor and a 32K ROM and 8K RAM are featured. Both ROM and RAM can be expanded by adding up to 16K.

A built-in automatic beacon can be set to transmit a message at designated intervals determined by the operator. Any station can act as a digital repeater and up to eight such "linking" stations are allowed which greatly expands the operator's range.

For complete information and a free catalog of Heath products, write Heath Company, Dept. 150-525, Benton Harbor, Michigan 49022. (In Canada write Heath Company, 1020 Islington Avenue, Suite 3100, Toronto, Ontario, M8Z 5Z3.)

Circle #319 on Reader Service Card.

flea market

RATES Noncommercial ads 10¢ per word; commercial ads 60¢ per word **both payable in advance**. No cash discounts or agency commissions allowed.

HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing) on a space available basis only. Repeat insertions of hamfest ads pay the non-commercial rate.

COPY No special layout or arrangements available. Material should be typewritten or clearly printed (**not** all capitals) and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** cannot check each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue.

DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

CX7 REPAIRS. 415-549-9210.

LETTERING and text for QSL pasteup and label usage. Stamp brings circular. Labelcraft, PO Box 412, W. Sand Lake, NY 12196.

HOUSE READY FOR HAM. Beautiful mountain top *all* brick 4 bedroom, 3 full bath home and radio site on one acre landscaped, private, with huge inground pool. Located 55 miles from Washington, DC and 60 miles from Baltimore, easy commute by Amtrack or van pools. Extremely energy efficient brick home, full basement and other extra rooms, radio room. All cable, poles, antennas, fire alarm system and intercom system stay. Must see to fully appreciate. Call or write to receive free written details of all features, $250,000. (301) 371-7521. John Murray, Box 290, Jefferson, Maryland 21755.

ELECTRON TUBES — Radio & TV types 75% off list price. Huge inventory. Also industrial types. Send for free catalog today or call Toll Free (800) 221-5802. Box HC, Transleteronice, Inc., 1365 - 39th Street, Brooklyn, NY 11218. See our display ad in this issue.

CUSTOM MADE embroidered patches. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept. 301, 4202 N. Drake, Chicago, IL 60618.

Foreign Subscription Agents for Ham Radio Magazine

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RECLAIM SILVER from electronics scrap! Write RALTEC, 25884F Highland, Cleveland, OH 44143.

WANTED: Manual and schematics for military V.L.F. receiver R-1134/WRR-3. Tom Adams, K9TA, PO Box 5541, Madison, WI 53705.

NEW CW Word Copying Course. Easy to learn. Complete OSO-Trainer™ Course. Send $16.95 (Indiana add 85¢). Money-back satisfaction guarantee (less $2.00 shipping and handling). Business-size SASE gets free information. AVC Innovations, Inc., Department HC, PO Box 20491, Indianapolis, IN 46220.

COMPUTERIZE Your IC-720 or R-70. Keyboard frequency entry, 64 memories, scanning (frequency, Memory, and mode). Many more features. Requires no interface, just cable, directions included. For Commodore 64. Cassette $14.75, disk $16.75, pdd. Cables also available. David Oliver, W9ODK, Rt. 2, Box 246, Shevlin, MN 56676.

WINDMILL & TOWER: *1978 Dunlite 2000 watt windmill, 110'* Rohn 45G tower with new concrete anchors, 110V - 220 amp battery bank, converter, loading switch. Complete system only $4,995.00. (518) 623-9951. Box 196, HCR-01, Warrensburg, NY 12885.

TRS-80 Model I/III/IV owners. HF antenna design program calculated dimensions for dipole, Yagi, and quad antennas. $14.95 (cassette) + $2.00 s&h to Cynwyn, Dept. H, 4791 Broadway, Suite 2F, New York, NH 10034.

MILITARY RADIOS: CPRC-26 Manpack Radio (described in March 1985 Ham Radio). Transceives 46-54 MHz, with battery box, antenna, crystal, handset: $22.50 apiece, $42.50/pair, good condition. R-390A Receiver, 5-32 MHz all modes, 4 mechanical filters, meters sealed (government removed, operation unaffected): $175 complete/checked; spair parts unit (80% complete, missing PTO/IF): $65. Info SASE. CPRC-26 add $4/unit shipping. R-390A shipping charges collect. Baytronics, Dept. HR, Box 591, Sandusky, Ohio 44870. 419-627-0460 evenings.

CLIMBING BELTS. OSHA certified nylon web belt with adjustable lanyard. $54.95 plus $3 UPS. John Limbach, K8NN, 7397 S. Platte Canyon Drive, Littleton, CO 80123.

MINT ROBOT 400. First sync modification. Instruction book and original box. I ship. $275 firm. Jim Valentino, KW2W, PO Box 438, Mastic Beach, NY 11951.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

CHASSIS and cabinet kits. SASE K3IWK.

IMRA, International Mission Radio Association, helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, NY 10538.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 09224 (213) 774-1255.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. P.O. Box RY, Cardiff, CA 92007.

OLD RADIO transcription discs wanted. Any size, speed. W7FIZ, Box 724 HR, Redmond, WA 98073-0724.

Coming Events

ACTIVITIES

"Places to go..."

OHIO: 43rd annual Findlay Hamfest sponsored by the Findlay Radio Club at the Hancock County Fairgounds. Sunday, September 8, 6:30 AM to 5 PM. Advance tickets $3.00 by September 1. At the door $4.00. Tables $6.00 each. Outdoor flea market spaces $3.00 each. Talk in on 147.75/15. For more information write Findlay Radio Club, PO Box 587, Findlay, Ohio 45839.

CALIFORNIA: The Sonoma County Radio Amateurs' third annual Ham Radio flea market Saturday, September 21, 8 AM to 2 PM, Sebastopol Community Center, 390 Morris St., Sebastopol. Admission and parking free. Tables $6/door, $5/advance. Vendor setup 7 AM. Radio clinic, exhibits, refreshments. Noon auction. Talk in on 146.13/73. For tickets and information: SCRA, Box 116, Santa Rosa, CA 95402.

WYOMING: (Laramie-Cheyenne) The 6th annual High Plains Ham Roundup, September 6, 7, 8, Medicine Bow National Forest, Yellow Pine Campground, 14 miles east of Laramie on I-80. Sponsored by the Shy-Wy ARC, University of Wyoming ARC and Northern Colorado ARC. Saturday potluck supper — bring your favorite dishes. Swapfest, packet radio, musical entertainment and campfire sing-a-long. All hams and families welcome. No registration fees, small forest service charge for campers. Talk in on 22/82 and 25/85. For information: KØHRS, 2204 Vassar Ave., Fort Collins, CO 80525.

PENNSYLVANIA: York Hamfest, September 21 and 22, York Fairgrounds, Rt. 74. Seminars, tailgating, displays, banquet and FCC exams on Saturday. Tailgating and displays on Sunday. Registration $2/Saturday and $4/Sunday. $5/both days. XYL's and Jr. ops under 12 free. Saturday evening banquet $10.00 advance only. Tailgating $4 per day; $6 both days. Indoor tables $5 and up per day. Vendors setup 6 AM. Registration 8 AM both days. Ladies' activities. Special motel rates and overnight camping on grounds. Write York Hamfest, Box W, Dover, PA 17315.

TEXAS: Houston Com-Vention '85, September 20-22, Stouffer Greenway Plaza Hotel, SW freeway and Edloe St., 5 miles SW of downtown Houston. Friday night registration 5:00 PM, hospitality suites and HARC auction. Saturday 8 AM to 5 PM. Sunday 9 AM to 3 PM. Indoor flea market, exhibits, ladies' luncheon and forums, free parking. Saturday night Texas BBQ dinner. For information (713) 333-1466.

NEW YORK: Hall of Science Amateur Radio Club Hamfest, Hall of Science parking lot, Flushing Meadow Park, 47-01 111th Street, Corona, Queens. Sunday, September 8. Raindate September 15. 9 AM to 4 PM. Donation: buyers $3.00, sellers $5.00 per space. Talk in on 144.250 simplex link, 223.6000 repeat, 445.225 repeat. For information: John Powers, KA2AHJ, (718) 847-8007. Arnie Schiffman, WB2YXB, (718) 343-0172.

NEW YORK: Long Island Hamfair, Sunday, September 22, New York Institute of Technology, Rt. 25A, Northern Blvd., Old Westbury. General admission $3. Spouse, children, sweethearts free. Exhibitors $5 per car space. No reservations. Talk in on 146.85. Hank Wener, WB2ALW (516) 484-4322 or Bob Reed, WB2DIN (516) 221-8116 evenings.

OHIO: Lima Hamfest, October 13, Allen County Fairgrounds, Rts. 309 and 117. Advance tickets $3.00. Door $3.50. Tables $6.00; half tables $3.50. For reservations SASE and check to NOARC, PO Box 211, Lima, OH 45801. Amateur exams, Novice through Extra. Send completed FCC 610 with check for $4.00 payable to ARRL/VEC and photocopy of current license by September 13 to Amateur Exams, NC8F, PO Box 211, Lima, Ohio 45802. SASE required.

NEW YORK: The 1985 Ham-O-Rama, September 14, Niagara Falls International Convention Center, Niagara Falls. General admission $3.50/advance before August 24, $5.00/gate. Canadian money accepted at par for advance general admission tickets. Computer and equipment displays, tech programs, indoor/outdoor flea market, FCC exams Novice through Extra. Outside flea market $5.00, inside $15.00. Talk in on 146.31/91 and 146.52 simplex. For information: Nelson Oldfield, 126 Greenaway Blvd., Cheektowaga, NY 14225.

1985 BLOSSOMLAND BLAST, Sunday, October 6, 1985. Write "BLAST", PO Box 175, St. Joseph, MI 49085.

KENTUCKY: The 1985 ARRL National Convention, October 4, 5 and 6, Kentucky Fair and Exposition Center, Louisville. Exhibitors and flea market all air conditioned. ARRL forum, packet radio, AMSAT, FCC National weather service, ladies' programs and much more. Admission $5.00 advance; $6.00 at door. 12 and under free. For information contact The Greater Louisville Hamfest Assn., PO Box 34444, Louisville, KY 40232. (502) 368-6657.

OHIO: The annual Cincinnati Hamfest, Sunday, September 15, Stricker's Grove, Rt. 128, one mile west of Venice (Ross). Exhibits, food and refreshments, flea market, music, hidden transmitter hunt, awards and sensational air show. Admission and registration $5.00. For information: Lillian Abbott, K8CKI, 317 Greenwell Road, Cincinnati, OH 45238.

MICHIGAN: The L'Anse Creuse ARC presents the 13th annual Swap and Shop, September 15, L'Anse Creuse High School, Mt. Clemens. 0800-1500. Plenty of food and parking. Trunk sales $4.00 per space. Inside tables $8.00 each. Tickets $2.00 at door, $1.00 advance. Talk in on 147.69/.09 and 146.52. For tickets and table reservations: Maurice Schietecatte, N8CEO, 15835 Touraine Ct., Mt. Clemens, MI 48044. (313) 286-1843.

VIRGINIA: ARRL Virginia State Convention and 10th annual Amateur Radio-Computer Fair, Saturday and Sunday, September 21 and 22, rain or shine, Virginia Beach Pavilion. 9 AM to 5 PM. Advance tickets both days $5.00. $6.00/door. Flea market tables $5.00 one day, $8.00 both days. Displays, forums, computer equipment, ARRL license upgrading exams. For information and tickets: Jim Harrison, N4NV, 1234 Little Bay Avenue, Norfolk, VA 23503. (804) 587-1695.

MICHIGAN: The Central Michigan ARC and Lansing Civil Defense Repeater Association are sponsoring Ham Fair '85 at the National Guard Armory, 2500 S. Washington Avenue, Lansing, Sunday, October 13, 8 AM to 3 PM. Admission $3.00. Tables 75¢ per foot by reservation only. FCC exams at 1 PM registration by September 13. For information and reservations: Rowena Elrod, KA8OBS, 111 Lancelot Place, Lansing, MI 48906. (517) 482-9650.

NEW HAMPSHIRE: The Connecticut Valley FM Association's 9th annual Hamfest and Flea Market, September 29, 9 AM to 5 PM rain or shine, King Ridge Ski Area, Sutton. General admission $2.00. Dealers and flea market set up $3.00. Food available. Overnight camping for SC units (no hookups). Talk in on 146.76 or 146.52 simplex.

CONNECTICUT: The Candlewood Amateur Radio Association (CARA) annual Flea Market, Sunday, September 15, Edmond Town Hall, Main St., Newtown, 10 AM to 4 PM. Dealers 9 AM. Admission $2. Tables $7. Tailgating $5. Talk in on 147.72/12 or 52 simplex. For table reservations send check or MO to CARA, PO Box 143, Bethel, CT 06801. For information: Gene Marino, W1IDH, Valley View Rd., Newtown, CT 06470. (203) 426-8852.

ILLINOIS: The 10th annual New Berlin Hamfest sponsored by the Sangamon County Fair Association, Sunday, September 22, rain or shine, Sangamon County Fairgrounds. 7 AM to 3 PM. Admission and flea market setup free. Food and drink available. Talk in on 146.52 and 146.88. For information: Al Swettman, K9QFR, Box 2, Pleasant Plains, IL 62677 (217) 626-1634.

OHIO: Cleveland Hamfest & Computer Show, Sunday, September 22, Cuyahoga County Fairgrounds, Berea. General admission 8 AM to 5 PM, $3.50 at gate, $3.00 advance. Under 12 free. Special NASA displays, vendors, speakers, Packet Radio, FCC and more. Non-ham and ladies' activities. For tickets or information: Cleveland Hamfest Assn., PO Box 93077, Cleveland, OH 44101. Walk-in Amateur Radio license exams starting 9 AM. Bring original and copy of current license. Check for $4 payable to ARRL/VEC. For information: Dave Willemin, AI8M, 331 Courtland, Elyria, OH 44035. Ph: 324-4574. Saturday night Hamfest banquet. For reservations: Barbara Ernest, N8DAD, 327-3914.

CONNECTICUT: Annual Natchaug Amateur Radio Association Flea Market, Sunday, September 22, 9 AM to 4 PM, Elk's Home, 198 Pleasant Street, Willimantic. Advance reserved tables inside/outside $5.00 ea. At door $7.00 ea. Admission $2.00. Under 16 free. Free parking. Talk in on 52 direct or 147.30/.90. For informaion: Ed Sadeski, 49 Circle Drive, Willimantic, CT 06226. (203) 456-7029 after 4 PM.

GEORGIA: DXPO 1985, September 27, 28 and 29, Lanier Plaza Hotel, I-85 and Monroe Drive NE, Atlanta. Reservations: Grover Meinert, KC8BX, 720 Starlight Lane NE, Atlanta, GA 30342. Registration $49.50.

PENNSYLVANIA: Pack Rats (Mt. Airy VHF ARC) 9th annual Mid-Atlantic VHF Conference, Saturday, October 5, Warrington Motor Lodge, Rt. 611, Warrington. 14th annual Hamarama, Sunday, October 6, Bucks County Drive-In Theatre, Rt. 611, Warrington. Flea market admission $5.00. Selling spaces $8.00 each. Gates open 6 AM, rain or shine. Bring own tables. Advance registration for Conference $4.00. Send to Hamarama '85, PO Box 311, Southampton, PA 18966 or Lee A. Cohen, K3MXM (215) 634-4942.

NORTH CAROLINA: 1985 National QCWA Convention, September 26, 27 and 28, Hyatt Winston-Salem. Meetings, forums and interesting local tours.

OREGON: The Walla Walla Valley Radio Amateur Clubs' 39th annual Hamfest, September 21 and 22, Community Building, Milton-Freewater. Free registration. For flyers write Pat Stewart, W7GVC, 1404 Ruth Street, Walla Walla, WA 99362.

GEORGIA: The 12th annual Lanierland ARC Hamfest, September 22, 9 AM Gainesville Holiday Inn. Free tables and inside display area for dealers with advance registration. Large parking lot for flea market. Doors open 8 AM for dealer set up. Talk in Novice through Extra exams starting 9 AM. Talk in on 146.07/.67. For information/reservations: Paul Watkins, W4FDK, Route 11, Box 536, Gainesville, GA 30501 (404) 536-8280.

NEW YORK: The Elmira Amateur Radio Association's 10th annual International Hamfest, September 28, Chemung County Fairgrounds. Outdoor flea market, dealer display, breakfast and lunch available on premises. Gates open 6 AM to 5 PM. Tickets at gate or advance from Steve Zolksky, 118 East 8th Street, Elmira Heights, NY 14903.

NEW MEXICO: Northern New Mexico ARC's annual Hamfest, September 28 and 29, Camp Stoney, 8 miles east of Santa Fe. Saturday AM — ARRL/VEC exams. Free overnight camping, no hookups. Sunday, 8 AM to 3 PM, tailgate flea market, dealers, programs. Admission $3.75 adults; $1.75 kids includes lunch. Talk in on .52 and local repeaters. For information: SASE to NNMARC, Rt. 3, Box 95-15, Santa Fe, NM 87501.

ILLINOIS: Peoria Area Amateur Radio Club's Superfest '85, September 21 and 22, Exposition Gardens, W. Northmoor Rd., Peoria. Admission $3.00 advance, $4.00 gate. Children under 12 free. Amateur radio and computer displays, huge flea market, FCC exams Saturday. Free bus to Northwoods Mall on Sunday. Full camping facilities on grounds. Saturday night get-together at Heritage House Smorgasboard, 8209 N. Mt. Hawley Rd., Peoria. For information: SASE to Superfest '85, PO Box 3461, Peoria, IL 61614.

MARYLAND: The Columbia Amateur Radio Assocition's 9th annual Hamfest, Sunday, October 6, 8 AM to 3:30 PM, Howard County Fairgrounds, 15 miles west of Baltimore. Admission $3.00. Spouse and kids free. Outdoor tailgating $5.00. Tables $6.00. Indoor tailgating (payments by Sept. 30) $6.00. Food available. Talk in on 147.735/147.135, 146.52/146.52. For reservations and information: Mike Vore, W3CCV. 9098 Lampskin Lane, Columbia, MD 21045. (301) 992-4953.

NEW YORK: Yonkers Electronics Fair and Giant Flea Market, Sunday, October 6, 9 AM to 4 PM, Yonkers Municipal Parking Garage, Nepperhan Avenue & New Main Street. Rain or shine. *Refreshments, free parking, unlimited free coffee all* day. Giant auction 2 PM. Admission $3.00. Children under 12 free. Sellers $7.00 per parking space, admits one. Bring tables. For further information: (914) 969-1053.

CONNECTICUT: CQ Radio Club of Torrington will hold its annual Flea Market, Saturday, October 5, 9 AM to 3 PM, East Albert Street Recreation Building. Dealers $7.00 per table. Tailgaters $5.00. Admission $1.00. Talk in on 146.955. For information: Donald D. Taylor, KA1GKJ, PO Box 455, Watertown, CT 06795.

NEW JERSEY: The South Jersey Radio Association, the oldest radio club in continuous operation in the US (1916) will hold its 37th annual Hamfest, Sunday, September 15, Pennsauken High School, Hylton Road, Pennsauken. Table and tailgating sales in parking lot. Refreshments and food available. Gates open 8 AM. Tickets $2.50 advance, $3.00 at door. Tailgating $5.00 per space plus admission ticket. Talk in 145.29/144.69. Contact Fred Holler, W2EKB, 348 Bortons Mill Rd., Cherry Hill, NJ 08034. (609) 795-0577.

NEW HAMPSHIRE: Hosstraders annual Fall tailgate swapfest, Saturday, October 5, Deerfield Fairgrounds. Donation $2 per person sellers included. Profits benefit Shriners' Boston Burns Center. Our May Swapfest gave $6,960. Friday night camping at Nominal fee after 4 PM. Talk in .52 and 146.40-147.00. For information SASE to Norm, WA1IVB, RFD Box 57, West Baldwin, ME 04091.

NEW YORK: The Radio Amateurs of Greater Syracuse 30th RAGS Hamfest, Saturday, October 5, New York State Fairgrounds off Rt. 690. 9 AM to 6 PM. Flea market setup 7:30 AM. Flea market tables $6. Outdoor tailgating $3. Exhibitors and dealers. VE exams by pre-registration. Speakers, ARRL forum, free parking. Admission $3. Talk in on 31/91 and 90/30. For information Viv Douglas, WA2PUU or Ed Swatiowski, WA2URK, PO Box 88, Liverpool, NY 13088.

RHODE ISLAND: The RI Amateur FM Repeater Service will hold its annual Fall Flea Market and Auction, Saturday, September 22, American Legion Fairmont Post 85, 870 River St., Woonsocket. Starts 9 AM. Spaces $5.00 each. Some other spaces available under pavilion. Auction noon to 5 PM. Admission free. Food and beverages available. Talk in on 34/94 and 52 simplex. For information: Richard Fairweather, K1KYI, 127 Sherman Farm Rd., Harrisville, RI 02830. (401) 568-3468.

MASSACHUSETTS: The Hampden County Radio Association will be giving exams for all classes of Amateur Radio licenses, Saturday, October 18, Hampden-Wilbraham Reg. High School, 621 Main Street, Wilbraham at 9 AM. Get Form 610 by writing to the club. For upgrading, bring original and copy of current license, check for $4 payable to ARRL/VEC. Send completed applications to Yorke Phillips, 235 Ames Rd., Wilbraham, MA 01036 or sign up that day. Sign up time 8:30 AM. Amateur radio classes will be held in the Fall. Send post card with name, address and phone number to Hampden County Radio Association, PO Box 482, West Springfield, MA 01090.

OPERATING EVENTS

"Things to do..."

Riding Radio Operators — Amateur Radio Motorcycle Club Net meets every Thursday night at 0300 UTC at 3888 kHz standard time and 7237.5 kHz daylight saving time. An eastern USA group meets one hour earlier at 3888 kHz year-round. Send business SASE to AG0N, Gary McDuffie, Rt. 1, Box 464, Bayard, NE 69334 and ask for net information.

September 21: Celebrate Connecticut's 350th anniversary with Connecticut DX Association and the Newington Amateur Radio League. Listen for KW1V operating from the grounds of the State Capitol in Hartford. For a special QSL card and official Connecticut Tourism map send business SASE with 39¢ to Dave Rose, KW1V, 13 Long Crossing Road, East Hampton, CT 06423.

September 18: The Little Brown Jug special event held by the Delaware Amateur Radio Association, W8QLS, at the Delaware County Fair. For a Delara QSL card SASE to W8QLS, Staff Stafford, 5987 Dublin Road, Delaware, OH 43015.

September 14: Members of the Utica-Shelby Emergency Communications Association (USECA) Utica, Michigan, will operate KA8KTV to celebrate the first air-to-ground public telephone service inaugurated between Chicago and Detroit 1957. For a certificate send large SASE to USECA, PO Box 291, Utica, MI 48087.

September 13: The Michigan Technological University ARC will operate special event station W8YY to celebrate the University's Centennial and the Club's Golden Anniversary. Help support young Amateur Radio students. For a commemorative certificate send QSL and large SASE to Debbie Parmer, c/o W8YY, W. Wadsworth Hall, MTU, Houghton, MI 49931.

September 10-14: The Southern Counties ARA will operate K2BR during the Miss America Pageant, Atlantic City, NJ. SASE via SCARA, Box 121, Linwood, NJ 08221.

September 29: Fall Classic Radio Exchange, 2000 UTC, Sunday to 0300 UTC, Monday. Object to restore, operate and enjoy older equipment with like-minded hams. Exchange name, RST, QTH, receiver and transmitter type. Same station may be worked with different equipment combinations, each band, each mode. CW call "CQ CX", phone "CQ Exchange". Send logs and SASE to Stu Stephens, K8SJ, 1407 Hollyrood Road, Sandusky, OH 44870.

September 14: The West Alabama Amateur Radio Society (WAARS) will operate special event station WD4DAT from the campus of the University of Alabama to commemorate the "Greatest College Football Coach in History" Paul 'Bear' Bryant. 1300Z to 2400Z. For a handsome commemorative certificate of the event send $1 and large SASE to WAARS, PO Box 1741, Tuscaloosa, AL 35403.

September 17: The Eagle Rock ARC of Idaho Falls will commemorate the 198th anniversary of the adoption of the Constitution of the United States with special event stations KX7C and N07B. For a special commemorative QSL SASE to Eagle Rock ARC.

September 21: The Paul Bunyan Wireless Association will operate a special event station from the site of the Paul Bunyan Festival near Brainerd, MN. For a commemorative QSL SASE to Paul Bunyan Wireless Assoc., PO Box 354, Pequot Lakes, MN 56472.

September 21: Minnesota QSO Party sponsored by the Paul Bunyan Wireless Association, 1700Z Saturday to 1700Z Sunday. Special awards for most contacts on phone, CW, mixed and QRP. Send logs to George Carleton, AD0S, PO Box 43, Merrifield, MN 56465.

September 12: The Columbiana County ARC will operate N8DKX to commemorate the annual Johnny Appleseed Festival. For a certificate send large SASE and QSL to N8DKX, 6008 Camp Blvd., Lisbon, OH 44432.

the ham notebook

solving a difficult TVI problem

Just when I thought I had the neighborhood TVI-proofed a new neighbor called to tell me my station was causing TVI to his cable service.

After many fruitless attempts to help him solve the problem, I told him that his situation was, in my opinion, due to the shoddy, run-of-the-mill converter that the cable company had provided. He didn't believe me and continued to write letters of complaint to the FCC.

In due time two FCC field engineers from the New York City office came to my place, dragged a truckload of instruments into the house, made a number of tests, gave my station a clean bill of health, and told the neighbor the TVI problem was *his* baby. Again I told the neighbor to get a better converter.

When the cable company did nothing about the problem, the neighbor bought a completely shielded state-of-the-art, crystal-controlled converter that cleared up most of the TVI.

We concluded that the residual interference may have been caused by reradiation from the neighbor's "Genie" automatic door units.

A call to the manufacturer was met not only with a sympathetic reception but also with quick action. The manufacturer's representative was familiar with the problem, which originated in the sequencer boards in the units. She said that even though the old units were out of warranty, the company would send new replacement boards at no charge provided the old boards were sent back. Installing the new boards (Alliance, Inc. Part No. 21277R) removed the last trace of TVI.

John Labaj, W2YW

THE GUERRI REPORT

Ernie Guerri W6MGI

RF power supplies achieve high efficiency

We have long come to expect the power supply to be the one part of any system that is heavy and hot. Typical linear power supplies have power densities of about 0.25 watt/inch³, and are about 35 percent efficient at converting AC line power to usable DC. In recent years, however, switching supplies have improved the situation by converting the AC line to square waves at about 20 kHz. This higher frequency permits switching schemes using pulse width modulation as a regulation mechanism, which in turn keeps the input-output differential across the regulator quite small. The resulting supplies have power densities of about 1 watt/inch³, and efficiencies of more than 60 percent. But this improvement is not without a price. Switching supplies generate considerable noise, and consequently require more extensive filtering and shielding than linear supplies. Even so, their efficiency is excellent, and switching supplies are now the prevalent type in most Amateur equipment.

Recently, National Semiconductor has made a significant advance in switching technology by developing a power module (No. HS9151) that operates at 1 MHz. This integrated package operates directly from the 120 volt AC line and can deliver 5 volts at 1 amperes with almost 80 percent efficiency. Its small size gives it a power density of over 5 watts/inch³. As would be expected, appropriate filtering is mandatory if you don't want to hear 1 MHz all over the spectrum. However, the high switching frequency makes filtering/shielding very manageable. This development surely signals the beginning of the end for bulky power supply components and frees valuable space in the new equipment designs for increased functionality.

receiver technology makes new strides

No matter what band we favor, part of our success or failure depends on the quality of the receiver we have working for us. Modern technology gives us the option to have nearly perfect receiving capability — if we choose. Sadly, much of today's excellent technology is not implemented in the latest production receivers.

It's *nearly* impossible to find a "professional" grade communications receiver whose front-end doesn't fall apart at the seams with input signals of only −10 dBm, and many do much worse. Yet we've had balanced JFET front-end designs with 3rd order intercepts of over +20 dBm for nearly 20 years! The incremental cost for incorporating these front-ends would be a few dollars in production. On the plus side, nearly all new VHF/UHF receivers incorporate GaAs FET input stages, which give good noise figures and can handle large, complex input signals.

Intermediate frequency and detector sections suffer similarly. The availability of monolithic crystal and ceramic filters has resulted in uniformly good performance at IF frequencies up to 45 MHz. However, few receivers offer multiple filters and all-mode detectors, even as options. Most of the necessary features are available in one or two ICs that could be added by the user if appropriate provisions were made during manufacture. Perhaps receiver designers could take a lesson from the computer people who leave empty card slots for expansion, or unpopulated sockets for memory expansion.

Thanks to the TV industry, frequency synthesis up to about 1 GHz is now relatively easy to achieve at modest cost. The Yaesu FRG-9600 is the first general coverage VHF/UHF receiver to use these low cost component assemblies in a versatile, reasonably priced package. These same techniques could be used to make high quality up-conversion (1st IF above the highest tuned frequency) receivers that would be image-free up through 450-500 MHz.

Just as ideas have been borrowed from the TV industry, we should soon see broader incorporation of computer technology into key receiver functions. The fact that most current equipment is microprocessor controlled makes the inclusion of an RS-232 port fairly easy, for example. Some manufacturers offer this feature now — but it (or something like it) will become almost a necessity as time goes by. Many of the advances now taking shape in the field of telecommunications are keyed to equipment that is computer controlled and highly adaptive.

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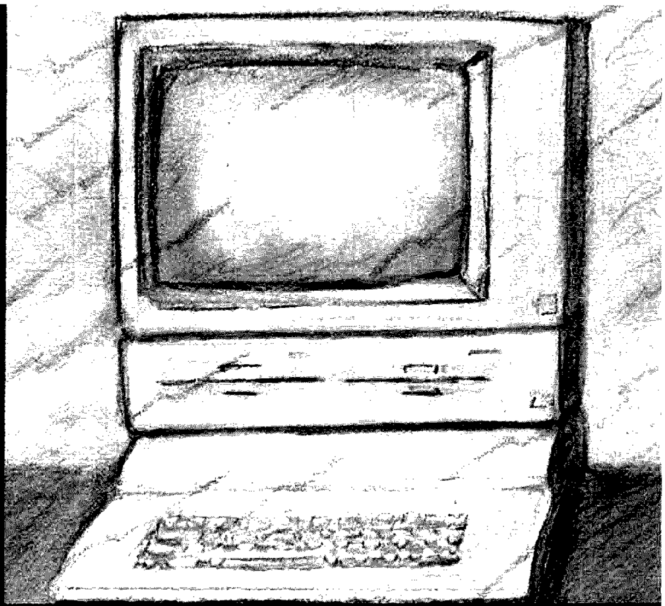
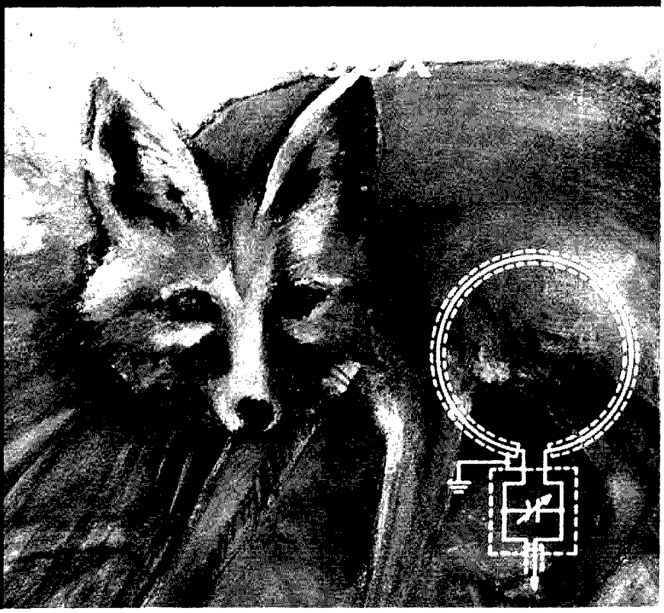
focus on communications technology





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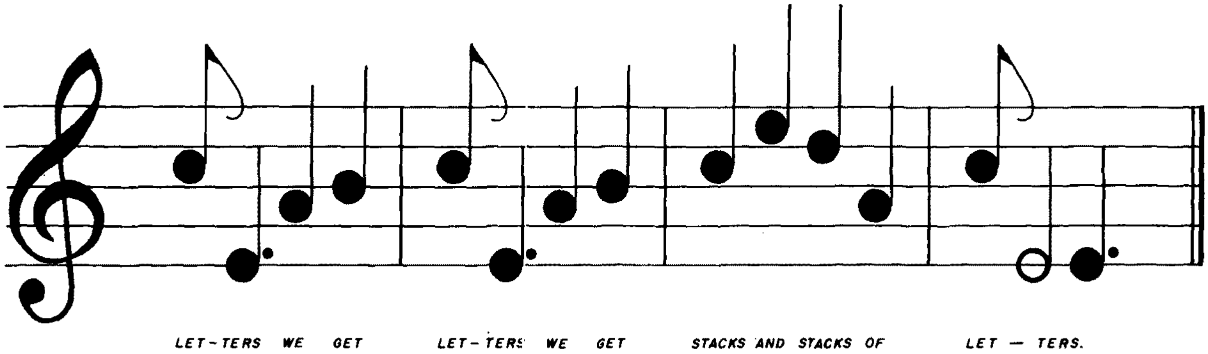


Dear Richie . . . Except for the name, the song's the same. I remember spending many an evening with my family, watching Perry Como on the Dumont. Towards the end of the program, mailmen would mysteriously appear, dump several large sacks of letters on the floor, and then disappear again, leaving Perry Como alone with those thousands of letters.

Sometimes at *ham radio* I feel that I'm in Mr. Como's situation. Every day we receive short letters and long letters from readers, asking or telling about simple subjects, easily grasped, and more complex subjects that send me to the quietude of my library. Club newsletters, new product announcements, DX news, and other written communications add to the pile.

Mind you, I'm not complaining. Quite honestly, I love to receive mail. As I'm sure I've mentioned many a time, reading material is like food to me — nay, like the breath of life itself. However (sorry, no "buts"), I've found that there *is* a limit to the number of hours in a day (it took me forty years to figure this out) . . . and though one should, in the name of efficiency, "prioritize," I must confess to a particular weakness: I believe that if someone's taken the time to write to the editor of *ham radio*, I want to take the time and care to respond. The reply might be shorter than you might like it to be — and it may take a long time in coming — but come it will.

There are a few things you can do to help make sure that you'll get an answer in a timely manner when you write. Here are a few suggestions:

DOUBLE SPACE your letter, regardless of whether you type, print, or knock it out on your word processor. Doing this makes your letter much easier to read and leaves room for me to scribble notes in between your lines.

Limit your comments to a SINGLE SUBJECT. Send as many letters as you like, but limit each one to one topic only. Sometimes several readers will comment on the same thing — for example, an error in a formula (although, of course, we rarely make mistakes). I can research your question and get an answer to you more expeditiously this way. (The DXers among you will appreciate this. When you've worked that rare one on several bands, don't you send separate cards for each one to make it easier on the QSL manager or the poor guy filling out those thousands of cards?)

Keep the LENGTH of you letter to a maximum of 400 words if you can. I understand that this may be difficult in some cases, but try. Make a game out of getting to the point quickly. (Now if only I could learn to do that!)

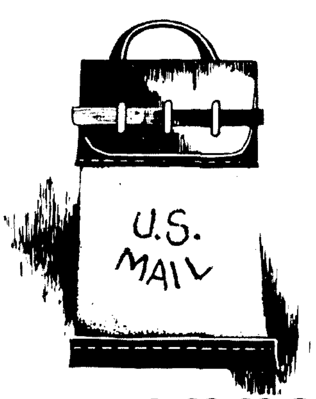
Clearly indicate if you'd like us not to PUBLISH your letter. I can't guarantee that we'll print your letter (if that's what you'd like), but I promise that if you say "do not print," your letter will not be printed. (If that's the case, please feel free to broach any subject. I don't embarrass easily.) One fine point: unless you tell us that you don't want your letter published, we'll assume we have your permission to publish it.

Don't be afraid that what you have to say may not be important, or that your letter might be too short (Grace à Dieu!). I started this editorial with very little to say, and look at it now.

Remember, that if nothing else, *ham radio* is a conduit of your thoughts and interests. Its content is gathered and published to meet your needs. The more readers we hear from the more accurate our understanding of your interests will be.

To the hardy few who've read this far, please write. And be patient. A response from this office *will* come.

Rich Rosen, K2RR
Editor-in-Chief



comments

more Novice privileges

Dear HR:

Back when the original Novice license was created, it included phone privileges on the high end of 2 meters. This was dropped because the new Novices were talking their 1-year, non-renewable tickets away. Now that the Novice ticket is a 10-year, renewable affair, maybe some thought should be given to letting the Novices back on 2 meters. . . . Interference the Novice may cause is easier to track down on 2 and does not propagate over the entire planet. Loading 2 meters with Novices might even force the development of modern commercial repeater techniques to increase the capacity of the voice repeater systems as well as development of the other VHF bands. . . .

The new license would allow Novices the same operating privileges as the Technicians, so they could operate packet, RTTY, CW, SSB, FM, or anything else that strikes their fancy and is legal.

This new Novice license would mean that Novices could work on public service and disaster communications while their interest is high (as was yours when you first got your license, remember?) and they're motivated to get involved. And they could also work on HF when they wanted to. The Techs will scream since many really aren't living up to the purpose of their license — the development of all the VHF spectrum — but are instead just living on 2 meters.

It would be nice if the ARRL would make holding at least one Novice class per year a requirement of affiliated clubs (forgiving schools — it's tough enough for those). Maybe also have a minimum passing rate of, say, 30 percent or three new Novices per year. Something like that, anyway. They want growth — this will help.

William E. Newkirk, WB9IVR
Melbourne, Florida

can anyone help?

Dear HR:

I am 31 years old, a General Class ham, and handicapped with multiple sclerosis. I am unable to work, so my wife supports us and our two little girls, ages 6 and 8. On our very limited income the purchase of new or used equipment is impossible.

Maybe there is someone in your readership who has some equipment they haven't used in a while, but is still in working condition, that they would be willing to donate. I need a good transceiver (solid-state or tube) or transmitter and receiver, power supply, and VHF/UHF radios, and any station accessories, such as microphone, watt/SWR meter, etc.

I can assure you that any equipment anyone donates will be put to good use and appreciated very much. Thank you for your time and effort on my behalf. God Bless you.

John T. Statham, N5HTQ
1506 Sheila Drive
McComb, Mississippi 39648
(601) 684-9558

service note

Although R.L. Drake no longer manufactures Amateur Radio products, service is still available. We're at 540 Richard Street, Miamisburg, Ohio 45342 (513 866-3211).

William A. Frost, WD8DFP
Service Manager
R.L. Drake Company

presstop de W9JUV

33 CM HAS BECOME THE NEWEST U.S. AMATEUR BAND, open for use 0001Z September 28, 1985, about the time you read this. All modes, for Technician class and above, are permitted in the 902-928 MHz slot, though Amateurs may not interfere with other band users (principally Industrial, Scientific, and Medical). Due to government needs, the band is not available in Colorado, Wyoming, or within the White Sands (NM) Missile Range; 33 cm Amateur stations within 150 miles of the Range borders are also limited to 150 watts PEP.

First Extensive Use Of 33 CM Is Likely To Be On FM, using 903-905 MHz rigs developed for use by the Japanese Personal Radio Service. A few are already in the U.S., and quite a few are believed to be in inventory in Japan since the number of Japanese users has not met the makers' expectations. W9JUV checked out a pair late last year on an FCC experimental license, and the Amateurs who had a chance to operate with them found their sophisticated selective calling and other features fascinating though absorption by foliage severely restricted their ground-level range. Unfortunately their frequency scheme conflicts with that developed by ARRL's VUAC, which places weak signal, beacon, linear translator and digital users in the 903-905 MHz portion of the new band and FM simplex at 906-907 MHz.

On The Same Date U.S. Amateurs Lost 420-430 MHz Above "line A," an imaginary line that runs across the country 50 miles south of the Canadian border. This provides a buffer zone for Canadian 420-430 MHz Land-Mobile activity, by a 1982 U.S.-Canadian agreement.

A NATIONAL ORGANIZATION OF VOLUNTEER EXAMINER COORDINATORS was probably the most significant result of the FCC's VEC meeting in Gettysburg August 9. 16 VECs, essentially all of the most active, were represented at the all-day session in which the FCC, for all intents and purposes, finished passing the Amateur examination ball to the Amateur VECs.

Though The FCC's Day-Long Meeting Was Itself Rated "Highly Successful" by all parties, it was at an informal post-session rehash Friday night that the seeds were planted for the national VEC group. Tentatively named the "Coalition of Amateur Radio Examiners" (CARE), the new group is considering such things as mutual accreditation of each other's Volunteer Examiners, developing a common examination pool with cooperative printing and stock-piling, and cooperation in scheduling of exams. Membership in CARE will be open to any FCC-accredited VEC and to any individual Volunteer Examiner even if the VE's VEC is not a CARE member.

Development Of The CARE Organization Is Continuing Rapidly; W9JUG at DeVry can provide additional information for those interested. A national VEC net, to discuss the VEC program in general and CARE in particular, now meets Sundays on 14173 kHz at 1700Z.

EXTENSIVE AMATEUR OPERATION FROM SPACE IS PLANNED for the Space Shuttle's Flight 61A, scheduled for liftoff October 16. The European crew includes Amateurs PE1LFO, DB2KM, and DD6CF, who'll operate on both 2 meters and 70 cm with antennas mounted on the Shuttle's surface instead of using the makeshift window antenna of previous flights.

Operation On The 10 Or 15 Meter Bands Is Also Being Considered, to provide new and useful information on HF propagation through the ionosphere. At press time it appears these operators will be permitted much more on-the-air time than any previous astronauts.

A U.S.-JAPANESE RECIPROCAL LICENSING AGREEMENT HAS BEEN SIGNED, according to the August 9 issue of Japan Times. Under U.S. reciprocal licensing rules, Japanese "no-code" license holders will have the same operating privileges here as they have in Japan—above 30 MHz. Japanese will not be the first "no-coders" to operate in the U.S., however, as a number of the other 65 countries with whom we have such agreements also have a no-code license. Since the Amateur population of Japan is so large, with so many holding a no-code license, a noticeable influx of Japanese reciprocal license holders can be expected.

8N1AAA Through 8N1XZZ Is The Callsign Block Reserved For U.S. Amateurs licensed in Japan. The new agreement, the result of extended negotiation (the U.S. is the first nation to establish a reciprocal agreement with Japan), should become effective in September.

CALIFORNIA'S PROPOSED BILL OUTLAWING 800-MHZ SCANNERS has been substantially tempered thanks to the efforts of attorney N6AHU and others. California Senate Bill 1431 still prohibits the "malicious" use of 800 MHz equipment designed specifically to eavesdrop on cellular radio, but includes exemptions for Amateurs and scanner buffs.

A Similar Threat Has Appeared On The National Scene, however, as a "discussion draft" of a U.S. House of Representatives bill that would broadly extend present restrictions against wiretapping to all forms of electronic communications.

Telephone Company Concern Over Privacy Of Cellular Communications is believed to have spawned this new bill, tentatively titled the "Electronic Surveillance Act of 1985," as was the case with the California legislation.

USE OF 432 MHZ AS A RELAY FREQUENCY FOR COMMERCIAL TV VIDEO has been requested in a request for waiver filed with the FCC by a Lake Havasu (AZ) low-power TV station owner. Comments on the TV station owner's request (which he claims has local Amateur support) are due October 10; Replies are due October 28. Refer to PRB-2; include four copies plus original

JORDANIAN AMATEURS WILL USE THE JY50 PREFIX November 7-21 in celebration of the 50th birthday of King Hussein, JY1. Extensive activity on 160 through 10 meters, plus OSCAR, is promised, with five JY50 QSO (10 in Europe) good for a special award.

computer-aided audio filter design

No adjustments necessary — just build it and it works

Audio filters are a simple, inexpensive way of realizing very high selectivity without any modifications to the transceiver (or receiver). A number of commercial audio filters are available, and many articles provide designs.

But I wanted an audio filter with specific features, including a very well-defined audio passband for CW applications; single-chip design (to minimize wiring); battery operation, with the lowest possible current drain to maximize battery life; and small size, for headphone-only operation. I also wanted to design and build this filter in just a weekend or two, and have fun doing it.

Although the "weekend or two" evolved into about two months of evenings and weekends, the project turned out to be both entertaining and educational.

Because I knew that two-pole bandpass or low-pass filters are very easy to design, I expected the project to be brief. The design equations are, after all, available in many reference books,[1] and I've listed them in **fig. 1** for convenient reference. Given the desired center frequency, *Q*, and gain, the correct Rs and Cs are easily calculated, (**fig. 1**). But a good audio filter requires that a number of two-pole filters be cascaded. So the question arises, what should the center frequency, *Q*, and gain of each two-pole section be? Should only bandpass sections be used, or should low-pass sections also be included? In other words, it's easy to design one stage, but it's not at all clear how to design three or four stages and know what the overall bandpass characteristics will be.

I approached this problem by writing a program for the Apple II+ that allows the user to enter four sets of *Q*, *H* (= gain), and *FØ* (= center frequency or corner frequency). Each filter may be either a bandpass or low-pass type. The program then evaluates the response of up to four filters over a user-designated band, in 10 Hz increments. The combined response of the four filters is then plotted on the screen using the high resolution graphics mode. This allows a detailed examination of the overall filter bandpass curve before it's built.

trial-and-error

This method, then, is essentially a computer-aided trial-and-error process. Four sets of parameters are selected, and the computer plots the response. If one side of the plot is "sagging," the gain, *Q*, or center frequency can be changed and the new response plotted. It isn't very long before one gets a feel of what's going to happen when a given parameter is changed.

By trying various combinations of *Q*, *H*, and *FØ* it takes surprisingly little time to obtain a filter that has a very flat top, and very steep skirts. Typically, it takes about ten tries; each takes less than 5 minutes. This means that a filter can be designed in about an hour. **Figure 2** shows the bandpass characteristics of a three-stage design that was done using the above process.

One might ask whether filters synthesized using this technique are the best of all possible filters that can be designed using up to four stages. This question may be addressed by comparing the results obtained using this method with the response of an ideal filter having an absolutely flat passband and infinitely steep skirts. It seems to me that passband flatness of better than 1/4 dB is pointless because you can't hear it — 1/4 dB is not perceptible to the human ear. For practical purposes, therefore, this filter is as good as the best obtainable as far as passband flatness is concerned.

However, it is possible, and even probable, that better skirts can be achieved using different filter parameters or perhaps four stages instead of only three. It becomes a question of how much time one is willing to put in to achieve the desired improvements.

By Dana F. Geiger, KE2J, 42A Sandy Hollow Road, Port Washington, New York 11050

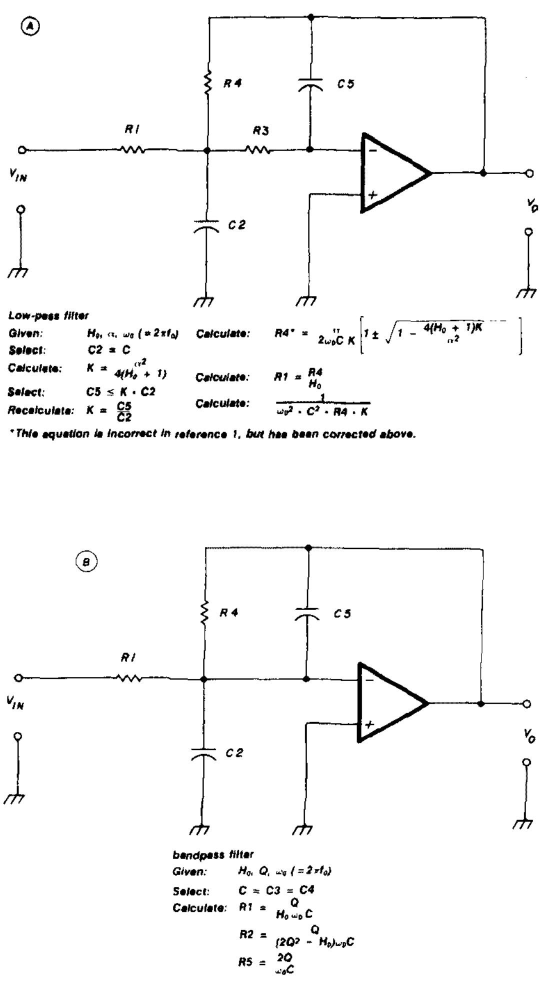


fig. 1. Design equations for single, two-pole (*A*) low-pass and (*B*) bandpass filter.

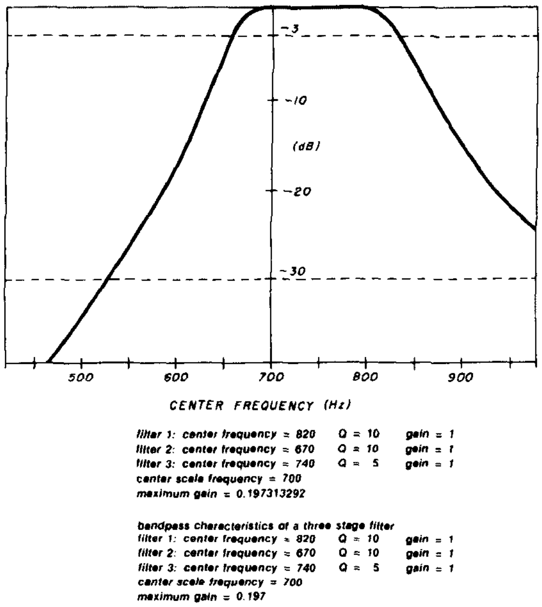


fig. 2. This shows plot of three cascaded bandpass filters (two poles each).

It's tempting to choose high *Q*s to improve the skirts. But this presents a potential problem because high *Q*s cause ringing. The rule of thumb I developed for choosing *Q* is:

$$Q = 10 \cdot F\emptyset$$

where $F\emptyset$ = *center frequency in kHz.*
Example: At $F\emptyset$ = 500 Hz, the maximum *Q* is 5. (Q = 10 • 1/2 kHz).

This rule of thumb is derived by calculating that at 40 WPM CW, a dot is about 25 ms long. This implies that the transient response of each stage of the filter should decay within 2 to 3 ms (= 1/10th of 25 ms), ensuring that the CW will not be stretched out by the decay time of the filter stages.

From the equation for a bandpass filter (see reference 1), it is known that the decay time constant is:

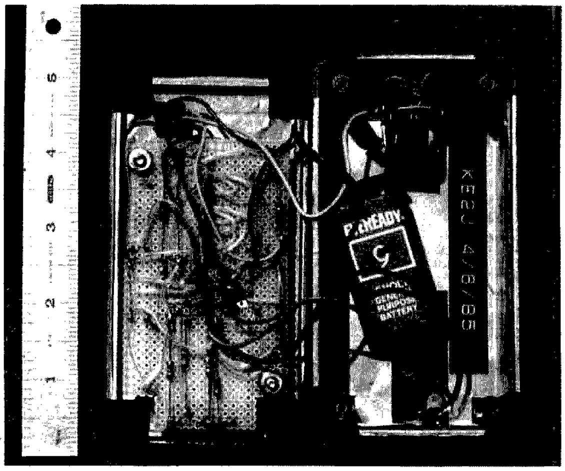
$$\text{decay time constant} = \frac{2 \cdot Q}{(2 \cdot \pi \cdot F\emptyset)} \qquad (1)$$

With the decay time constant selected at about 3 ms, the above rule of thumb follows. This is really a somewhat conservative rule, and serves only as a guide. In the above example, a *Q* of 10 is still not unreasonable, but a *Q* of 25 or 50 will create a music synthesizer, not an audio filter.

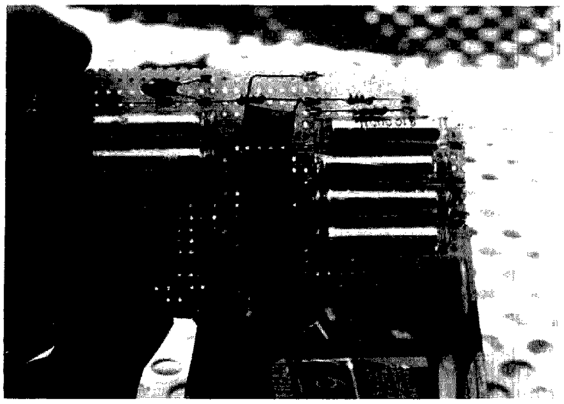
The 3 dB width of the filter was selected as somewhere between 150 and 200 Hz because it seemed to me that it would be difficult to tune in a signal with a substantially narrower filter bandwidth. A 200 Hz width represents only a very small rotation on the transceiver main tuning dial. Furthermore, there is a limit to how narrow the bandwidth can be made for a given CW speed. 200 Hz seems to be quite effective in practice.

After calculating the values of resistors and capacitors for the filter (using the design equations in **fig. 1**), it was not surprising that the calculated values were not standard 5 percent parts. However, the closest standard 5 percent values were substituted and the analysis program was then used to recalculate the filter parameters.

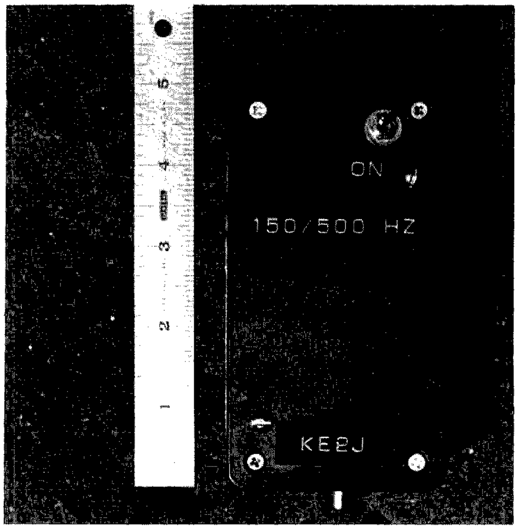
This resulted in filter parameters that differed slightly from the originally synthesized values. For example, the initial *Q* of 10 became 9.9. To be sure that the



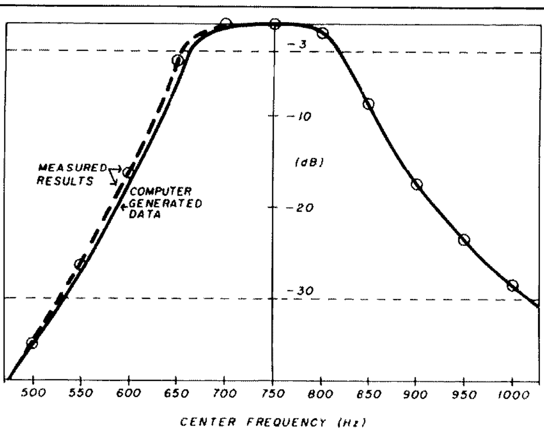
Simple construction techniques include point-to-point wiring on perfboard.



Heart of circuit is contained in centrally located 14-pin DIP.



Completed unit measures less than 3 × 5 inches (7.6 × 12.7 cm).



computer generated vs measured bandpass characteristics
filter 1: center frequency = 808 Q = 9.9 gain = 1.01
filter 2: center frequency = 671 Q = 9.9 gain = 0.98
filter 3: center frequency = 740 Q = 5.0 gain = 1.08
center scale frequency = 750
maximum gain = 0.246

measured results (see schematic)

frequency (Hz)	output volts	dB = 20 log (output volts / peak volts)
500	1.8 mV	−35.00
550	4.8 mV	−26.40
600	15.4 mV	−16.30
650	63.6 mV	−4.00
700	100.6 mV	−0.01
750	99.2 mV	−0.13
800	89.5 mV	−1.02
850	36.4 mV	−8.80
900	13.8 mV	−17.30
950	6.8 mV	−23.40
1000	3.8 mV	−28.50
716 (peak)	100.7 mV (max)	0 (reference)

maximum gain (at 716 Hz) $= \frac{100.7\ mV}{116.8\ mV} = 0.86$*

(V_{in} = 116.8 mV at all frequencies)

*circuit contains an amplifier — see schematic (fig. 4).

fig. 3. The "y" axis has been placed at 750 Hz by using the manual mode in the plotting program.

overall filter characteristics didn't change substantially, the transfer function was replotted using the program. If significant departures were observed, closer values of resistors were synthesized using parallel combinations of resistor pairs. The replotted bandpass curves, shown in **fig. 3**, can be compared to the original plots in **fig. 2** to demonstrate how the use of practical parts has changed the filter.

filter circuit

The filter circuit is shown in **fig. 4**. The nomenclature for the resistors and capacitors follows that in reference 1. The entire circuit draws 1.3 mA from the 9 volt battery, while delivering full volume into 8 ohm headphones.

The performance was measured using a digital voltmeter (true RMS), a digital counter, and a function generator. In **fig. 3**, the actual response is plotted over the computer generated response.

The nastiest problem was the presence of RFI in the headphones during transmit. W2CXK suggested lining the inside of the plastic box with copper foil. (A metal box is probably best for the enclosure.) It was also necessary to use RF filters at both the input and output to totally eliminate the RFI.

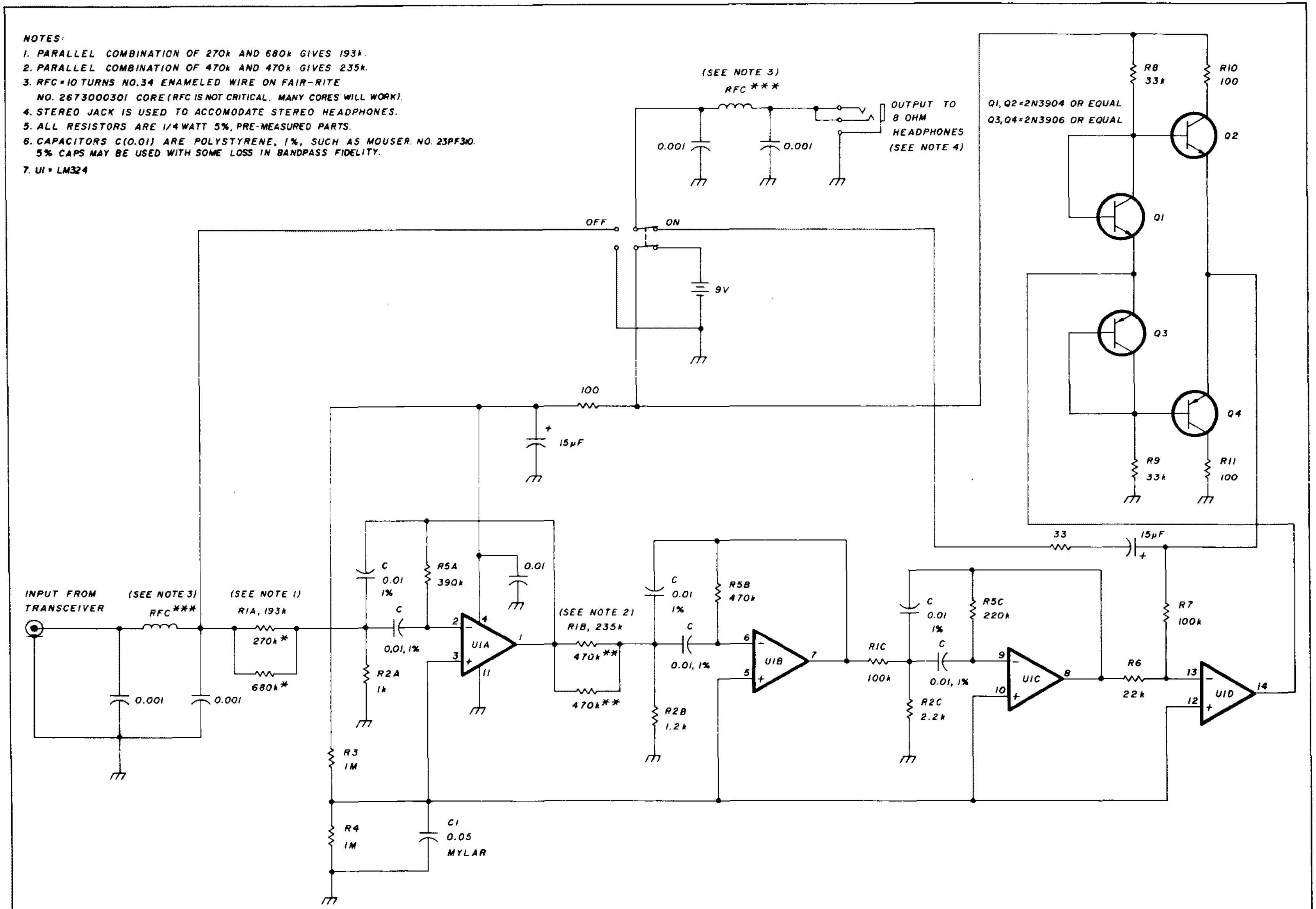


fig. 4. Schematic of audio filter.

```
10 X = FRE (0)
20 PRINT : PRINT
30 PI = 3.1415926
40 TEXT : HOME
50 DEF FN L(X) = LOG (X) / LOG (10)
60 REM LOG(X) IS THE NATURAL LOG IN THIS BASIC. LOG (10) IS THE NATURAL
    LOG OF 10 (ABOUT 2.3). L(X) IS THE LOG TO THE BASE 10, AS REQUIRED IN
    THE PROGRAM.
70 DEF FN H(W) = H0 / SQR (1 + (Q ^ 2) * (W / W0 - W0 / W) ^ 2)
80 DEF FN LP(W) = (H0 * W0 ^ 2) / SQR (W ^ 4 + (AL ^ 2 - 2) * (W ^ 2) *
    (W0 ^ 2) + (W0 ^ 4))
90 REM ******************************
100 REM ENTER FILTER CHARACTERISTICS
110 INPUT "TODAY'S DATE:";D$
120 INPUT "HOW MANY FILTER SECTIONS? (1,2,3 OR 4): ";FS
130 IF FS = 1 THEN T2 = 0:T3 = 0:T4 = 0: GOTO 180
140 IF FS = 2 THEN T3 = 0:T4 = 0: GOTO 180
150 IF FS = 3 THEN T4 = 0: GOTO 180
160 IF FS = 4 THEN 180
170 HOME : GOTO 120
180 PRINT "***** FILTER 1 *****"
190 INPUT "BANDPASS OR LOWPASS?(B OR L):";T$
200 IF T$ = "B" THEN INPUT "CENTER FREQUENCY=";F1: INPUT "Q=";Q1: INPUT
    "GAIN=";H1:T1 = 1: GOTO 230
210 IF T$ = "L" THEN INPUT "CORNER FREQUENCY=";F1: INPUT "ALPHA=";A1: INPUT
    "GAIN=";H1:T1 = 2: GOTO 230
220 HOME : GOTO 180
230 IF FS = 1 THEN GOTO 410
240 PRINT : PRINT "***** FILTER 2 *****"
250 INPUT "BANDPASS OR LOWPASS?(B OR L):";T$
260 IF T$ = "B" THEN INPUT "CENTER FREQUENCY=";F2: INPUT "Q=";Q2: INPUT
    "GAIN=";H2:T2 = 1: GOTO 290
270 IF T$ = "L" THEN INPUT "CORNER FREQUENCY=";F2: INPUT "ALPHA=";A2: INPUT
    "GAIN=";H2:T2 = 2: GOTO 290
280 GOTO 240
290 IF FS = 2 THEN GOTO 410
300 PRINT : PRINT "***** FILTER 3 *****"
310 INPUT "BANDPASS OR LOWPASS?(B OR L):";T$
320 IF T$ = "B" THEN INPUT "CENTER FREQUENCY=";F3: INPUT "Q=";Q3: INPUT
    "GAIN=";H3:T3 = 1: GOTO 350
330 IF T$ = "L" THEN INPUT "CORNER FREQUENCY=";F3: INPUT "ALPHA=";A3: INPUT
    "GAIN=";H3:T3 = 2: GOTO 350
340 GOTO 300
350 IF FS = 3 THEN GOTO 410
360 PRINT : PRINT "***** FILTER 4 *****"
370 INPUT "BANDPASS OR LOWPASS?(B OR L):";T$
380 IF T$ = "B" THEN INPUT "CENTER FREQUENCY=";F4: INPUT "Q=";Q4: INPUT
    "GAIN=";H4:T4 = 1: GOTO 410
390 IF T$ = "L" THEN INPUT "CORNER FREQUENCY=";F4: INPUT "ALPHA=";A4: INPUT
    "GAIN=";H4:T4 = 2: GOTO 410
400 GOTO 360
410 PRINT : PRINT
420 INPUT "STARTING FREQUENCY=";FB
430 INPUT "ENDING FREQUENCY=";FE
440 ARG = INT (FE - FB) / 10
450 IF ARG < = 0 THEN GOTO 420
460 DIM R(ARG)
470 PRINT : PRINT "CENTER FREQUENCY OF GRAPH"
480 PRINT
490 PRINT "ENTER 'M' FOR MANUAL"
500 PRINT "(ANY OTHER CHARACTER DEFAULTS"
510 PRINT " TO AUTOMATIC.)"
520 GET M$
530 IF M$ = "M" THEN INPUT "CENTER FREQ=";FC
540 REM ***********************************
550 FOR F = FB TO FE STEP 10
560 W = 2 * PI * F
570 Q = Q1:H0 = H1:W0 = F1 * 2 * PI:AL = A1
580 IF T1 = 1 THEN G1 = FN H(W): GOTO 600
590 IF T1 = 2 THEN G1 = FN LP(W): GOTO 600
600 IF T2 = 0 THEN G2 = 1: GOTO 640
610 Q = Q2:H0 = H2:W0 = F2 * 2 * PI:AL = A2
620 IF T2 = 1 THEN G2 = FN H(W): GOTO 640
630 IF T2 = 2 THEN G2 = FN LP(W): GOTO 640
640 IF T3 = 0 THEN G3 = 1: GOTO 680
650 Q = Q3:H0 = H3:W0 = F3 * 2 * PI:AL = A3
660 IF T3 = 1 THEN G3 = FN H(W): GOTO 680
670 IF T3 = 2 THEN G3 = FN LP(W): GOTO 680
680 IF T4 = 0 THEN G4 = 1: GOTO 720
690 Q = Q4:H0 = H4:W0 = F4 * 2 * PI:AL = A4
700 IF T4 = 1 THEN G4 = FN H(W): GOTO 720
710 IF T4 = 2 THEN G4 = FN LP(W)
720 R((F - FB) / 10) = G1 * G2 * G3 * G4
730 PRINT "R(";F;")=";R((F - FB) / 10)
740 NEXT F
750 REM ********************************
760 REM FIND THE LARGEST RESPONSE (LEAST ATTENUATION) IN THE BAND.
770 MAX = R(0)
780 FMAX = FB
790 FOR N = 0 TO ARG - 1
800 X = N + 1
810 IF R(X) > R(N) THEN MAX = R(X):FMAX = X * 10 + FB
820 PRINT N
830 NEXT N
840 REM CONVERT TO DB WITH REFERENCE TO THE LARGEST SIGNAL.
850 FOR N = 0 TO ARG
860 X = R(N) / MAX: REM NORMALIZE
870 R(N) = 20 * FN L(X)
880 PRINT N,R(N)
890 NEXT N
900 REM ****************************
910 REM PLOT AXES
920 HGR : HCOLOR= 3
930 HPLOT 140,0 TO 140,159: REM Y AXIS
940 HPLOT 0,0 TO 0,156: REM LEFT BORDER
950 HPLOT 279,0 TO 279,156: REM RIGHT BORDER
960 HPLOT 0,156 TO 279,156: REM X AXIS
970 HPLOT 0,0 TO 279,0: REM TOP BORDER
980 FOR N = 1 TO 27: REM X AXIS TICS
990 REM SCALE FACTOR IS 2HZ PER PIXEL. THEREFORE IN 140 PIXELS, 280 HZ I
    S DISPLAYED ON EITHER SIDE OF THE CENTER FREQUENCY
1000 REM A TIC EVERY 10 HZ
1010 HPLOT 140 + 5 * N,156 TO 140 + 5 * N,157
1020 HPLOT 140 - 5 * N,156 TO 140 - 5 * N,157
1021 REM : 3DB AND 30DB LINES
1022 HPLOT 140 + 5 * N,12
1023 HPLOT 140 - 5 * N,12
1024 HPLOT 140 + 5 * N,120
1025 HPLOT 140 - 5 * N,120
1030 NEXT N
1040 REM A LARGE TIC EVERY 50 HZ
1050 FOR N = 1 TO 5
1060 HPLOT 140 + 25 * N,0 TO 140 + 25 * N,2
1070 HPLOT 140 - 25 * N,0 TO 140 - 25 * N,2
1080 HPLOT 140 + 25 * N,156 TO 140 + 25 * N,159
1090 HPLOT 140 - 25 * N,156 TO 140 - 25 * N,159
1100 NEXT N
1110 FOR N = 0 TO 152 STEP 4
1120 HPLOT 139,N TO 141,N
1130 HPLOT 0,N TO 1,N
1140 HPLOT 278,N TO 279,N
1150 NEXT N
1160 REM Y AXIS TICS AND LINES
1210 HPLOT 138,40 TO 142,40
1220 HPLOT 0,40 TO 2,40
1230 HPLOT 277,40 TO 279,40
1240 HPLOT 138,80 TO 142,80
1250 HPLOT 0,80 TO 2,80
1260 HPLOT 277,80 TO 279,80
1261 HPLOT 138,120 TO 142,120
1262 HPLOT 0,120 TO 2,120
1263 HPLOT 277,120 TO 279,120
1270 REM X AXIS SCALE FACTOR IS 2HZ/PIXEL
1280 REM Y AXIS SCALE FACTOR IS 10DB/40 PIXELS
1290 REM *******************************
1300 REM PLOT THE RESPONSE
1310 FF = FMAX
1320 IF M$ = "M" THEN FF = FC
1330 FOR N = 0 TO 54
1340 X = 5 * N + 5
1350 F = FF - 270 + 10 * N
1360 IF F > FE THEN 1460
1370 IF F < FB THEN 1480
1380 Y = 4 * ABS (R((F - FB) / 10))
1385 REM FACTOR OF 4 BECAUSE THERE ARE 4 PIXELS PER DB. THIS ESTABLISHE
    S THE SCALE FACTOR ON THE SCREEN.
1390 IF Y > 159 THEN 1480
```

fig. 5. Plotting program listing.

fig. 5. (continued)

```
1400 YL = Y - 2:YU = Y + 2
1410 XL = X - 2:XU = X + 2
1420  IF YL < 0 THEN YL = 0
1430  IF YU > 159 THEN YU = 159
1440  IF XL < 0 THEN XL = 0
1450  IF XU > 279 THEN XU = 279
1460  HPLOT XL,Y TO XU,Y
1470  HPLOT X,YL TO X,YU
1400  NEXT N
1490  PRINT "CENTER FREQ=";FF,"GAIN=";MAX
1500  PRINT "PARAMETER HARDCOPY? (Y OR N)"
1510  GET Y$
1520  IF Y$ = "Y" THEN  GOSUB 1570
1530  IF Y$ = "N" THEN 1550
1540  GOTO 1500
1550  END
1560  REN ****************************
1570  REM  MAKES HARDCOPY OF PARAMETERS.
1580  PR# 1: REM  TURN ON PRINTER
1590  PRINT D$
1600 H0 = H1:Q = Q1:F = F1:AL = A1:T$ = "FILTER1:"
1610  IF T1 = 1 THEN  GOSUB 1810: REM  FOR BPF
1620  IF T1 = 2 THEN  GOSUB 1860: REM  FOR LPF
1630  IF T2 = 0 THEN 1670
1640 H0 = H2:Q = Q2:F = F2:AL = A2:T$ = "FILTER2:"
1650  IF T2 = 1 THEN  GOSUB 1810
1660  IF T2 = 2 THEN  GOSUB 1860
1670  IF T3 = 0 THEN 1710
1680 H0 = H3:Q = Q3:F = F3:AL = A3:T$ = "FILTER3:"
1690  IF T3 = 1 THEN  GOSUB 1810
1700  IF T3 = 2 THEN  GOSUB 1860
1710  IF T4 = 0 THEN 1760
1720 H0 = H4:Q = Q4:F = F4:AL = A4:T$ = "FILTER4:"
1730  IF T4 = 1 THEN  GOSUB 1810
1740  IF T4 = 2 THEN  GOSUB 1860
1750  PRINT
1760  PRINT "CENTER SCALE FREQUENCY=";FF
1770  PRINT "HAXIMUM GAIN=";HAX
1780  PR# 0: REM  PRINTER OFF
1790 X =  FRE (0)
1800  END
1810  REM  ***************************
1820  REM  PRINT ROUTINE FOR BPF'S
1830  PRINT
1840  PRINT T$;"  CENTER FREQ=";F;"  Q=";Q;"  GAIN=";H0
1850  RETURN
1860  REN ***************************
1870  REM  PRINT ROUTINE FOR LPF'S
1880  PRINT
1890  PRINT T$;"  CORNER FREQ=";F;"  ALPHA=";AL;"  GAIN=";H0
1900  RETURN
1910  REM ***************************
```

I have used this audio filter on the air with my FT101ZD, which has very good selectivity by itself. Yet the audio filter provides a very substantial improvement in reading the CW signals through QRM or QRN. It is important to have a means of switching the filter in and out during a normal QSO. Even a small amount of drift of either station will throw the signal out of the passband. It is then necessary to revert to wideband operation (that is, filter off) to find the other station again! This is accomplished with the on/off switch shown in the schematic. The signal is directly routed to the headphones in the "off" position

```
10  TEXT : HONE
20 X =  FRE (0)
30  PRINT : PRINT
40 PI = 3.1415926
50  PRINT "SELECT THE DESIRED FUNCTION BY NUMBER
60  PRINT
70  PRINT "I) DESIGN BANDPASS"
80  PRINT
90  PRINT "2) ANALYZE BANDPASS"
100  PRINT
110  PRINT "3) DESISN LONPASS"
120  PRINT
130  PRINT "4) ANALYZE LONPASS"
140  PRINT
150  PRINT "5) EXIT"
160  SET S
170  ON S GOTO 210,400,620,930,610
180  GOTO 10
190  REM *********************************
200  REM  BPF DESIGN
210  HOME
220  INPUT "GAIN=";H0
230  INPUT "Q=";Q
240  INPUT "CENTER FREQUENCY (HZ)=";F0
250 W0 = 2 * PI * F0
260  INPUT "GELECT C(UF)=";C
270 C = C * 10 ^  - 6
280 R1 = Q / (H0 * W0 * C)
290 R2 = Q / ((2 * Q ^ 2 - H0) * W0 * C)
300 R5 = 2 * Q / (N0 * C)
310  PRINT "R1=";R1
320  PRINT "R2=";R2
330  PRINT "R5=";R5
340  PRINT "ANOTHER DESIGN? (Y OR N)"
350  GET Y$
360  IF Y$ = "Y" THEN  GOTO 210
370  IF Y$ = "N" THEN  GOTO 10
380  GOTO 340
390  REM  ***************************
400  REM  BPF ANALYSIS
410  HOME
420  INPUT "R1(KOHM)=";R1
430 R1 = R1 * 10 ^ 3
440  INPUT "R2(KOHM)=";R2
450 R2 = R2 * 10 ^ 3
460  INPUT "R5(KOHM)=";R5
470 R5 = R5 * 10 ^ 3
480  INPUT "C(UF)=";C:C = C * 10 ^  - 6
490 H0 = 1 / (2 * (R1 / R5))
500 W0 =  SQR ((1 / (C * C * R5)) * ((1 / R1) + (1 / R2)))
510 A = 1 /  SQR ((1 / R1 + 1 / R2) * R5)
520 Q = 1 / (2 * A)
530  PRINT "SAIN=";H0
540  PRINT "Q=";Q
550  PRINT "F0=";W0 / (2 * PI)
560  PRINT "ANOTHER EVALUATION (Y OR N)?"
570  GET Y$
580  IF Y$ = "Y" THEN  PRINT : GOTO 420
590  IF Y$ = "N" THEN 10
600  GOTO 560
610  END
620  REH **************************
630  REM   LPF DESIGN
640  INPUT "LON FREQ GAIN (H0)=";H0
650  INPUT "ALPHA (<2)=";AL
660  INPUT "BREAKPOINT FREQ=";F
670 N0 = 2 * PI * F
680  INPUT "C2 (UF)=";C
690 K = (AL ^ 2) / (4 * (H0 + 1))
700  PRINT "CHOOSE C5 <=";K * C;" (UF)"
710  INPUT "C5 (UF)=";C5
```

fig. 6. Design/analysis listing **(continued on page 22).**

```
720  IF C5 > K * C THEN  PRINT "TRY AGAIN..": GOTO 700
730 K = C5 / C
740 C = C * 1E - 6
750 D =  SQR (1 - 4 * K * (H0 + 1) / AL ^ 2)
760 R4 = (AL / (2 * K * W0 * C)) * (1 + D)
770 S4 = (AL / (2 * K * W0 * C)) * (1 - D)
780 R3 = 1 / ((W0 ^ 2) * K * (C ^ 2) * R4)
790 S3 = 1 / ((W0 ^ 2) * K * (C ^ 2) * S4)
800 R1 = R4 / H0
810 S1 = S4 / H0
820  PRINT "R1=";( INT (1000 * R1)) / 1E6;" KONS"
830  PRINT "R3=";( INT (1000 * R3)) / 1E6;" KOHMS"
840  PRINT "R4=";( INT (1000 * R4)) / 1E6;" KOHMS"
850  PRINT "******** OR ********"
860  PRINT "R1'=";( INT (1000 * S1)) / 1E6;" KOHMS"
870  PRINT "R3'=";( INT (1000 * S3)) / 1E6;" KOHMS"
880  PRINT "R4'=";( INT (1000 * S4)) / 1E6;" KOHMS"
890  PRINT "ANOTHER DESIGN? (Y OR N)"
900  GET Y$
910  IF Y$ = "Y" THEN 620
920  GOTO 10
930  REM  ****************************
940  REM  LPF ANALYSIS
950  INPUT "R1 (KOHMS)=";R1
960  INPUT "R3 (KOHMS)=";R3
970  INPUT "R4 (KOHMS)=";R4
980  INPUT "C2 (UF)=";C2
990  INPUT "C5 (UF)=";C5
1000 R1 = R1 * 1E3:R3 = R3 * 1E3:R4 = R4 * 1E3
1010 C2 = C2 * 1E - 6:C5 = C5 * 1E - 6
1020 H0 = R4 / R1
1030 W0 =  SQR (1 / (R3 * R4 * C2 * C5))
1040 AL = ( SQR (C5 / C2)) * ( SQR (R3 / R4) +  SQR (R4 / R3) + (1 / R1) *
      SQR (R3 * R4))
1050  PRINT "GAIN=";H0
1060  PRINT "CORNER FREQ=";W0 / (2 * PI)
1070  PRINT "ALPHA=";AL
1080  PRINT "ANOTHER ANALYSIS? (Y OR N)"
1090  GET Y$
1100  IF Y$ = "Y" THEN 930
1110  GOTO 10
```

fig. 6. (continued).

the plotting program

The program proceeds in the following sequence:

- Information is requested by the program, and Q, gain, and center frequency are entered for up to four filter sections. The desired frequency range and placement of the "y" axis are also entered.
- The response is calculated over the frequency range in 10-Hz increments and stored in an array.
- The largest response is determined.
- Each element of the array is normalized and converted to dB with respect to the largest array element.
- The results are plotted on the high-resolution graphics screen. The "y" axis is marked in 1 dB steps, and 1 pixel is 1/4 dB. Therefore, the resolution is 1/4 dB. The "x" axis is marked in 10 Hz steps. Each pixel represents 2 Hz; hence the resolution is 2 Hz per pixel.
- The "y" axis is always in the middle of the screen, and there is always a span of 280 Hz on either side of the "y" axis. The program prints out the frequency corresponding to the "y" axis, which calibrates the graph.
- In the automatic mode, the graph is plotted so that the maximum response is always on the top of the "y" axis. This ensures that the most significant points will appear. In the manual mode, the "y" axis frequency is specified by the user. This allows examination of points outside the 560 Hz span, or placement of the "y" axis at the center of the response curve, even if it's not the peak.

A listing of the plotting program is provided in **fig. 5**, and the design/analysis program is listed in **fig. 6**. An Apple II+ disc (5-1/4 inch, DOS 3.3) with the program and design notes may be ordered from Electronics Unlimited, 42A Sandy Hollow Road, Port Washington, New York 11050. The price is $25 (postpaid).

reference

1. J.G. Graeme, G.E. Tobey, and L.P. Huelsman, PhD., *Operational Amplifiers — Design and Application*, McGraw Hill, 1971.

ham radio

the fox box
a direction-finding tool

Add this compact unit to your 2-meter rig and catch that fox — quickly

This is a circuit intended for 2-meter fox hunting. This inexpensive and easy-to-build device, nicknamed the "Fox Box," is a remote signal-strength meter and wide-range variable front-end attenuator that, when needed, quickly attaches to your 2-meter FM transceiver. Because this article is written for beginners as well as experienced hunters, I've also included a photo and construction details for a simple, low-cost, two-element 2-meter quad. I claim no originality for these circuits, except perhaps for some refinements that I have made after doing numerous modifications to friends' radios. Similar schemes have been circulatng among fox hunting groups for some time.

Let's consider the two basic parameters a fox hunter works with: directivity and signal strength. Directivity is provided by either a manually-orientated antenna array, or via the use of an automatic doppler-shift system.[1,2] In either case the radio's internal S-meter is used to indicate signal strength. Unfortunately the limited dynamic range, and compressed nonlinearity on strong signals, of most FM radio metering circuits requires the use of an external attenuator to limit the signal levels reaching the receiver. The Fox Box allows precise control of the receiver gain to assist with the antenna orientation.

After using the Fox Box for a short time the fox hunter develops a feel for distance based on experience with the meter readings and the amount of attenuation required. Fox hunts are won or lost in the last mile; hunters without good attenuators are often misled into looking for the fox far from the actual hiding place. They're still half a mile out and have full-scale readings regardless of where they point their antennas.

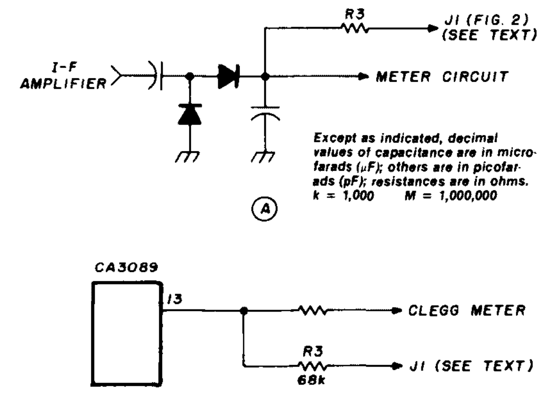


fig. 1. Representative 2 meter FM transceiver (meter) detector circuits: (*A*) commonly used by ICOM, Kenwood, Yaesu, KDK, and Azden. R3 is determined empirically; refer to *table 1* and text. (*B*) Clegg FM-DX uses the CA3089 FM system IC. 68 kilohms is about optimum for the Radio Shack 50 μA meter.

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a better way

Many hunters rely on external attenuators using resistive elements to reduce strong signals. But most external attenuators are at best cumbersome, and the minuscule S-meters adorning FM transceiver front panels are at best useless. LED bargraph displays fare no better, their resolution is coarse, and often the display will not show signals that are plainly audible. External attenuators suffer from other problems. Stray RF pickup (through the power leads and through leaks in the radio enclosure) limit the amount of external attenuation that may be used. Hunters, in the heat of competition, have forgotten about their inline attenuators and attempted transmissions, only to be rewarded with a wisp of smoke and charred resistors. The Fox Box eliminates these problems.

will it work on my radio?

The external metering and attenuator control box built for a Clegg FM-DX transceiver is shown in photo 1. I've also provided details for several popular transceivers (see fig. 1, table 1). This selection is not arbitrary — these radios best exemplify all of the variations encountered while modifying several other different transceivers.

table 1. Resistors (as specified in owner's manuals or in service manual schematic diagrams) that must be rerouted for the external attenuator modification.

transceiver	resistors
Azden PCS2000	R4, R5, R10
Clegg FM-DX	RX-8 feed
Clegg FM-28	RX-8 feed
ICOM IC-22S	R7, R13
KDK FM2025	R4, R13, R17*
Kenwood TR7950	R7, R15, R21
Kenwood TR7850	R5, R13
Kenwood TR7600/25	R41, R34, R35

*Denotes resistors involved in third stage.

The Fox Box will work on any 2-meter transceiver using MOSFET devices in the front-end circuits. It should work as well on JFET circuits, too, although this has not been tried. Some transceivers made in the 1960s and early 1970s used bipolar technology in the first RF stages; a different approach will have to be used in these vintage radios. Several likely methods of controlling bipolar front-end stages are noted in this section.

the internal attenuator

The internal attenuator circuit is a very useful modification. The Fox Box provides only a DC control level for setting the attenuation, and all RF remains contained within the radio (see figs. 2, 3, 4). Most FM transceivers use two dual-gate MOSFET devices in the front end — the first as the RF amplifier and followed by another in the first-mixer stage. Some radios have a third MOSFET, employed as a post-mixer amplifier, after the first crystal filter. When present, this stage may also be placed under attenuator control (more on these radios later). Figure 3 is a generic receiver front-end circuit, representative of the almost universal approach followed by VHF FM transceiver manufacturers.

A regulated voltage source, normally between 8 and 9 volts, powers these stages. By changing this voltage the receiver's gain is controllable over a very wide range. While this is a brute-force approach, it is also very effective. The attenuator modification consists of installing a MPS-A14 Darlington transistor in series with the supply voltage to these stages. Forward bias, supplied through a base-to-collector fixed resistor, keeps the transistor in full-conduction, thus permitting normal receiver operation with the Fox Box disconnected. When the Fox Box is in use its internal potentiometer, in conjunction with the fixed-value resistor, forms a variable voltage divider, the MOSFET's supply voltage can then be set from about 7 volts down to zero volts for full attenuation (fig. 2).

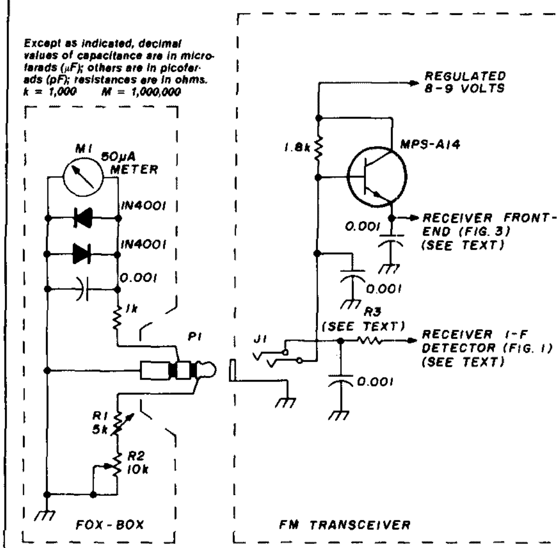
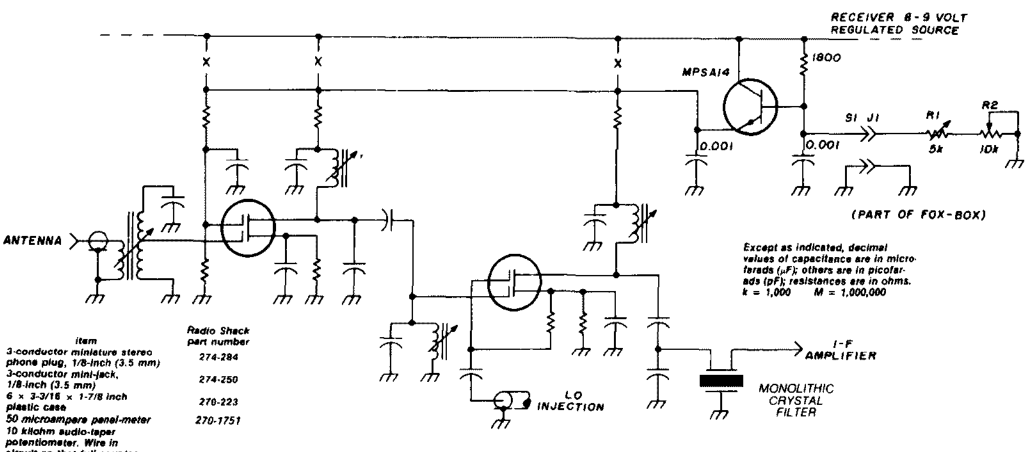


fig. 2. Fox-Box circuit includes remote box with auxiliary meter and attenuator (see *photo 1*) and FM transceiver interface. Diodes and a 1000-ohm current limiting resistor protect the meter from accidental damage. Trimpot R1 (5 kilohms) is optional, (see text). In the modified radio, resistor R3 sets the full-scale reading for meter M1. The active circuit components for the attenuator are mounted in the radio.



item	Radio Shack part number
3-conductor miniature stereo phone plug, 1/8-inch (3.5 mm)	274-284
3-conductor mini-jack, 1/8-inch (3.5 mm)	274-250
6 × 3-3/16 × 1-7/8 inch plastic case	270-223
50 microampere panel-meter	270-1751
10 kilohm audio-taper potentiometer. Wire in circuit so that full counter-clockwise (CCW) rotation is equal to minimum resistance: full attenuation	271-1721
5-kilohm trimpot (see text)	271-217

fig. 3. Typical 2-meter receiver front end includes internal attenuator installation details.

A few comments on attenuation systems for bipolar stages: one possibility involves shunting the base-bias resistors. The RF amplifier will probably be biased for class A operation, and the bias will most likely be developed across a resistive voltage divider. An external potentiometer may then be used to lower the forward bias, thus reducing the stage gain dramatically. A good rule of thumb is to use a resistance three to four times the value of the base-to-ground resistor. For example, if a 5.6 kilohm resistor is used, a 20 kilohm potentiometer will work well for the attenuator control. Normally the base-biasing resistors will be cold — that is, no RF will be on them. If they're in the RF path, suitable RF decoupling chokes and bypass capacitors will be needed. **Figure 4** shows suggested attenuator installations for bipolar RF stages.

the remote signal-strength meter

The Fox Box also contains the external S-meter (**photo 1**), an inexpensive, 2-1/2 inch (6.33 cm) movement, large enough so that small signal-strength variations are readily observed. With a sufficiently long cable attached, the Fox Box attenuator and meter permit quick antenna orientation while away from the mobile-radio installation. (My relative signal-strength metering circuit is shown in **fig. 2**.) Most FM transceiver signal-strength display circuits use two diodes in a voltage-doubler configuration to sample the RF levels at the IF stage output (**fig. 1A**). The Clegg FM-DX transceiver is an exception: it uses an RCA CA3089 chip for the IF amplifier, FM detector, and meter driver. Pin 13 of the CA3089 produces an increasing voltage proportional to the input signal. The Radio Shack meter is a 50-μA movement — a series current-limiting resistor (R3) is used between the meter and IF detector. About 68 kilohm is needed for the FM-DX (**figs. 1B**, and **2**). In all cases the resistor is fine tuned to produce a full-scale reading with a full saturation input signal. Sometimes the no-signal meter reading will idle above the zero mark; this is normal for some radios and results from detected noise produced by some high gain IF systems. Full-scale external meter deflection can be best set via the initial use of a 200-kilohm trimpot for R3 in transceivers that use a diode detector. Once the correct resistance value is determined, a fixed-value resistor may be substituted for the potentiometer. When the Fox Box is used the rig's internal signal strength readings may be lower than normal due to the loading effect of the external meter.

installing the Fox Box

Modifying radios for the internal attenuator requires some dexterity. The Clegg FM-DX is the easiest because its RF amplifier and mixer are on one circuit board; one lead *(the one with blue insulation) brings* the power into these circuits. Since the forward-bias for gate 2 of the MOSFETs must be controlled as well as the drain voltage, it's important to be sure that all of the resistors carrying power to these stages are lifted from the V_{CC} runs and rerouted to the MPS-A14 emitter. Only the resistor lead connected to the V_{CC} supply run is lifted. With the resistor standing vertically, a length of hookup wire is tack-soldered to the free lead. A length of heatshrink tubing over the resistor and hookup wire junction keeps things neat. The MPS-A14 can be mounted on a three-terminal solder-lug phenolic strip — install the strip where space permits.

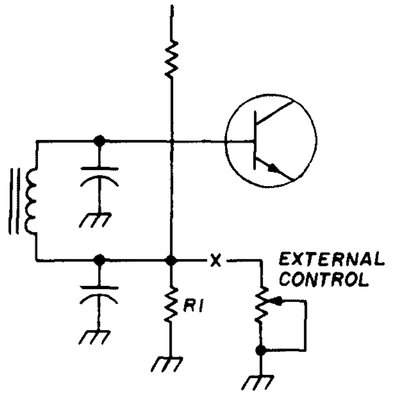


fig. 4. RF gain control scheme for tranceivers using bipolar transistor front ends. Unlike the MOSFET attenuator, no supply voltage variation is used; instead the base biasing voltage is shunted through an external potentiometer. The value for the potentiometer is typically four to five times higher than the base-to-ground biasing resistor (R1).

Sometimes only one resistor per stage need be rerouted — look for a single low-value resistor that provides voltage to a stage, such as in the KDK FM-2025, where resistor R4 (47 ohms) supplies power to RF amplifier Q1. Again, in the FM-2025, for mixer-stage Q2 only resistor R13 (220 ohms) is rerouted. Gate 2 bias is developed from a voltage divider (comprised of R4 and R15) located on the FET's source side of R13. **Table 1** references owner-manual schematics supplied with the Azden PCS-2000, Kenwood 7950, 7850, and 7625, KDK FM-2025, and Clegg FM-DX and lists the resistors required for each of these models. (If your radio is not one of those listed, studying **fig. 3** will help in determining the resistors involved in your transceiver. If you send me (at my home address) a good copy of your schematic, I'll circle the appropriate resistors — an SASE *must* accompany your request.)

inside the fox box: variations and adjustments

The schematic for the internal attenuator is shown in **fig. 2**. Part of the circuit involves two resistors: R2, the Fox-Box front panel control used for setting the desired attenuation level, and trimpot R1, which is used only in radios with three stages under attenuation control. The maximum attenuation will be more than required, R1 sets the maximum attenuation level. When operating three stages near maximum attenuation, the squelch may open. For some radios this is normal, because these stages may contribute a large portion of the total receiver gain and reducing their gain will affect noise operated squelches. There was one minor shortcoming with this attenuator. When approaching maximum attenuation, some users mentioned that the control became very nonlinear and touchy to adjust. In retrospect, I had been using linear potentiometers and found that audio-taper potentiometers proved the better choice. The potentiometers must be wired so that maximum attenuation occurs at the CCW position.

interconnecting the transceiver and the Fox Box

The Fox Box is connected to the radio via a three-conductor cable using a miniature 1/8-inch (3.5 mm) three-conductor "stereo" plug. One lead is the common-ground return for the attenuator and meter, and the remaining leads carry the meter and attenuator signals. Use enough cable to allow you to leave the car, Fox Box in hand, so you can move about while orienting the directional antenna. This is the real beauty of this device — it allows you to do some DFing *outside* the car. With the Fox Box disconnected, the radio reverts to normal premodification operation. Never connect or disconnect the adapter while the radio is on.

Some hams may have reservations about drilling the mounting hole for the 1/8-inch stereo jack. If the external speaker jack isn't used, and no future use is contemplated, the jack may be removed, or taped and left in the radio. Some radios, such as the Clegg FM-DX, have 9-pin accessory jacks, and spare pins are often available. Some deterioration of receiver dynamic-range may occur when using the attenuator. This has not been observed in actual use — but with the reduced mixer-stage voltage levels it is a possibility.

All necessary components can be obtained at electronic supply stores for under $20. Installation typically takes only three or four hours.*

using the Fox Box

The internal attenuator, at full attenuation, will give about a half-scale reading from a 25-watt mobile 10 feet (3.04 meters) away. In many hunts the fox has been hidden in a high-reflection area, effectively eliminating the doppler and Yagi competition. When this has happened I've been able to get in using just the attenuator and S-meter provided in the Fox Box — as the fox is talking, the meter indicates whether you're within a few thousand feet of its location, and whether approaching or leaving its location.

*If you're interested in this adapter, but feel ill-at-ease about tearing into your radio, I'm willing to do the work in my service shop at the going rate. Write, enclosing an SASE, and we'll work out the details.

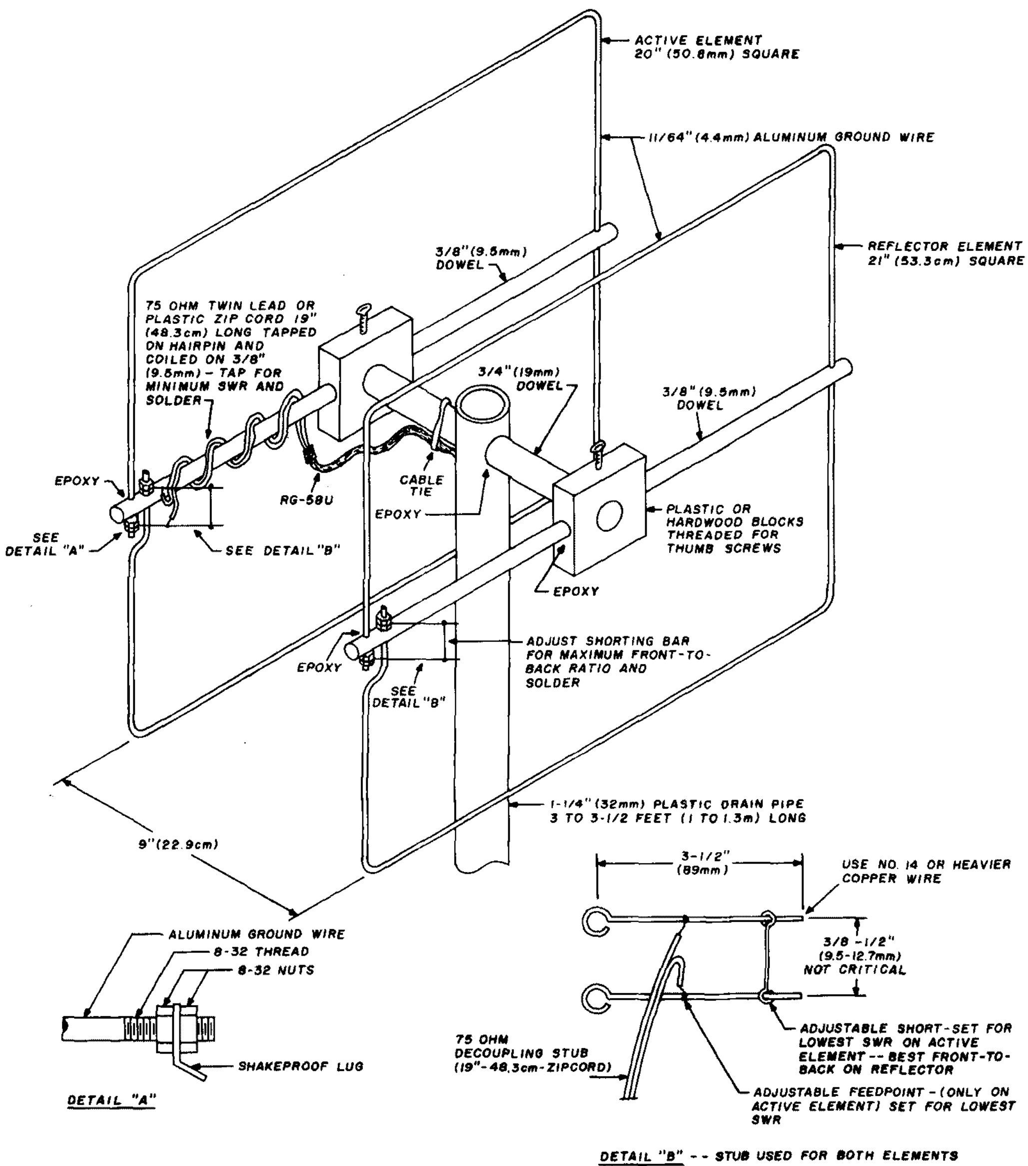


fig. 5. Light and durable quad for hidden transmitter hunting. A variety of materials may be used for the boom and masting — wooden doweling or PVC pipe are a few good choices. The elements are made from either aluminum ground or copper wire. Note the 75-ohm decoupling section wrapped on the driver boom. (Figure courtesy WA3TNO.)

K3TS two-element 2-meter quad antenna

As promised, here are the details for a simple DFing antenna. This design is the work of Tom Stewart, K3TS. A detailed mechanical drawing of the antenna is shown in **fig. 5**. Tom reported that the antenna did quite well while hilltopping and in mobile operation. Tom also designed the mobile antenna mount pictured in **fig. 6**. This mount allows the driver to steer the antenna while the car is moving. The lower (bottom) mount is inserted between the window glass and rubber sealer, while the upper mount is screwed (*using short screws!*) or otherwise affixed to the window frame. For installation in frameless windows, or "no-holes" mounting, the upper bracket is fastened to the end of a single rooftop carrier positioned directly above the door-mounted unit.

Portable antennas can take a beating during fox hunts. K3TS uses aluminum wire for his quad elements. This allows the antenna to "give," without damage, so that when a low branch or other obstruc-

tion bends the elements they may be easily reshaped. (Softdrawn copper antenna wire might prove better — it can be soldered, thereby avoiding the problems of making good electrical connections to aluminum wire.) The closed loop design results in little antenna interaction from the presence of the car's body. Using K3TS's design, one need only loosen the thumbscrews and rotate the antenna elements 90 degrees for horizontal polarization (feedpoint top or bottom).

Do not omit the all important 75-ohm decoupling section. This decoupler was first suggested by E.M. Brown, in *CQ* back in 1952. The front-to-back ratio will suffer if not used. Plastic insulated zipcord or speakerwire can be used in lieu of the 75-ohm twinlead. The 75-ohm twinlead is tightly wound, without overlaps, along the length of the element support dowel. The coax-to-twinlead junction is as close to the boom-to-element T-block as possible.

tuning the antenna

Begin the alignment by removing the reflector assembly from the boom. Tune the quad by adjusting the balun position on the feedpoint hairpin, while alternately adjusting the hairpin shorting-stub, for minimum SWR. (Don't expect a "perfect match" with this antenna; an SWR of 1.5:1 is acceptable.) The initial settings given in **fig. 5** are a good starting point. (Use miniature fleaclips on the balun and shorting-bar to aid in the initial positioning.) When the driven element is properly adjusted, the reflector is installed. The shorting stub on the reflector hairpin is then set for the best front-to-back ratio (or null). The antenna produces 6 dB forward gain with upwards of 25 dB of front-to-back ratio.

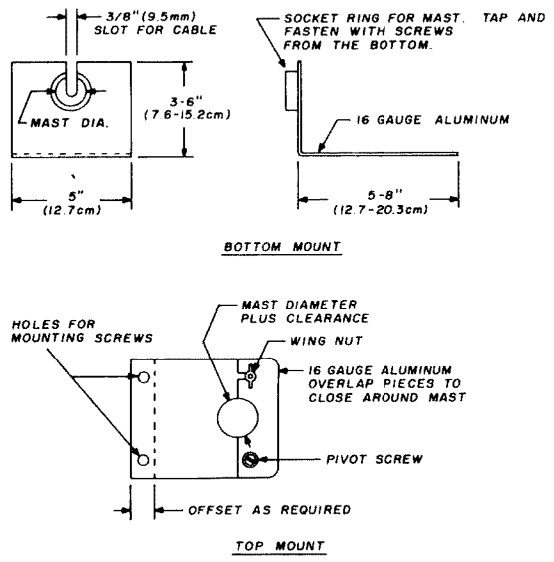


fig. 6. Simple mounting brackets, made from 16-gauge aluminum, permit operation of the quad while vehicle is in motion. (Figure courtesy WA3TNO.)

performance

The Fox Box is shown in **photo 1**. The antenna was built by Bob, KA1IQD — a closeup of it is shown in **photo 2**.

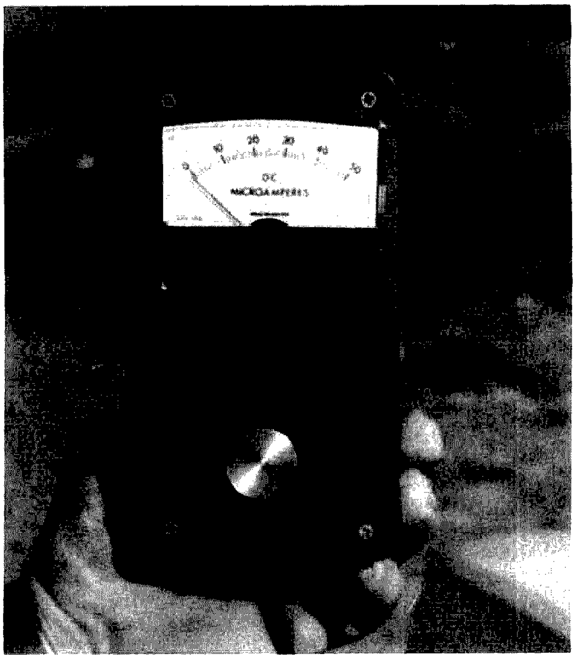


Photo 1. The K1ZJH Fox Box external metering and attenuator control box.

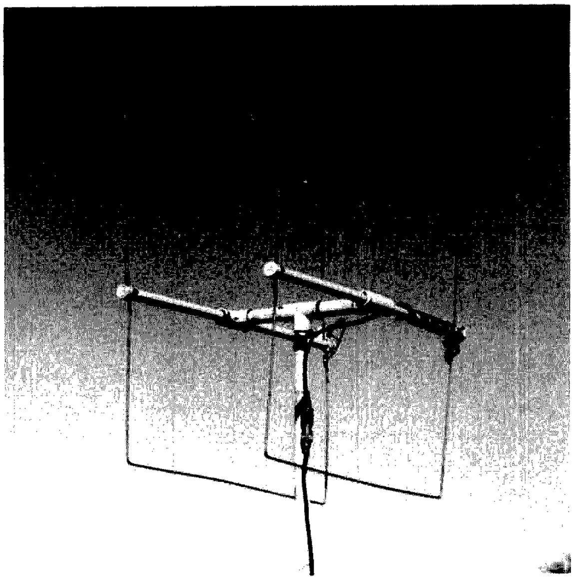


Photo 2. Closeup view of the K3TS quad as built by Bob, KA1IQD. Bob used PVC tubing and stainless steel wire for constructing his quad. Numerous copies of the original design have been successfully completed by area hams using various materials and construction techniques.

Although this is a good beginner's antenna, many experienced hunters swear by it. Don't be misled by its small size and simplicity. (Bob, KA1IQD, and I recently used his Fox Box and quad to locate a spurious signal jamming a local repeater input frequency on 147.600 MHz. The source of the signal turned out to be a mast-mounted TV antenna booster located several miles from the repeater site.) I was very impressed with the antenna and found it easy to use.

getting started

A few tips for the prospective fox hunters reading this: first, as with anything new, a little hands-on experience will help you to become a proficient DF'er. Make several practice runs using the attenuator before going on a hunt — this will give you a "feel" for its operation. To start, have a friend hide nearby while you and a partner use your DF'ing gear to locate him. Because the driver's only concern must be the safe, legal, and proper operation of the motor vehicle, a team of two people works out best; let your partner read the maps and interpret headings while underway.

references

1. Terrance Rogers, WA4BVY, "A DoppleScant," *QST*, May, 1978, page 24.
2. David Cunningham, W7BEP, "DF Breakthrough!" *73*, June, 1981, page 32.

ham radio

building and using phone patches

From simple to elegant, patches help make the connection

In telephone company parlance, a patch is any connection between a phone line and another communications device, whether it be a radio, a tape recorder, a data device (such as a modem), or even another phone line.

Radio Amateurs, on the other hand, tend to limit the meaning of "patch" to the connection of transmitters or receivers to the phone line for phone conversations. But there's more to it — Amateurs can and do use phone patches for purposes other than telephone conversations. One particularly effective application is for checking TVI and RFI complaints; simply set the transmitter on VOX, go to the site of the interference complaint, and then key your transmitter via the phone line. Doing this will indicate whether your transmitter is or is not the source of the problem. If it is, you can use this method to test the measures you've taken to correct the problem.

A phone line is, simply speaking, a 600-ohm balanced feed device — which also happens to be how professional audio can be described. Most modern Amateur transmitters have 600-ohm unbalanced inputs; most cassette recorders have a 600-ohm unbalanced input; the "tape" outputs on home stereos are also 600-ohm unbalanced. All this makes patching relatively simple. While there are various degrees of sophistication and complexity in patching, in an emergency, patches can be easily put together using readily available components. Before starting to build a patch, however, it might be helpful to read last month's article on understanding phone lines.[1]

the simple patch

The simplest way to patch a phone line to another piece of equipment is to use a couple of capacitors to block the phone line DC. While this simple approach will work in a pinch, it will tend to introduce hum to the line because of the unbalance introduced. The capacitors used should be nonpolar, at least $2\mu F$, and rated at 250 volts or better (see **fig. 1**).

To hold the line, the patch should provide a DC load by means of a resistor (R6) or by simply leaving a phone off the hook. The receiver output may need a DC load (R7) to prevent the output stage from "motorboating." Use two capacitors to maintain the balance.

With all patches hum can be lessened by reversing the phone wires. A well-made patch will have no discernible hum.

the basic phone patch

Because a phone line is balanced and carries DC as well as an AC signal, a patch should include a DC block, a balun, and a DC load *to hold the line*. The best component for doing this is a 600-ohm 1:1 transformer such as those used in professional audio and for coupling modem signals to the phone line, available from most electronics supply houses. Old telephone answering machines are also a good source of 600-ohm transformers. Some transformers are rated at 600:900 ohms or 900:900 ohms; these are also acceptable. Make sure that the transformer has a large enough core, because DC current will be flowing through it. (Some small-core transformers become saturated and distort the signal.)

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In section 68.304 of the FCC Part 68 regulations, it states that a coupling transformer should withstand a 60 Hz 1kV signal for one minute with less than 10 mA leakage. For casual use this may seem unimportant, but it provides good protection against any destructive high voltage that may come down the phone line, and into the Amateur's equipment. A 130 to 250 volt Metal Oxide Varistor (MOV) across the phone line will provide further protection if needed.

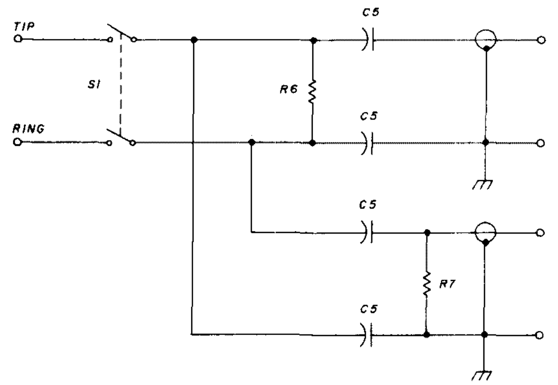

fig. 1. Simple phone patch.

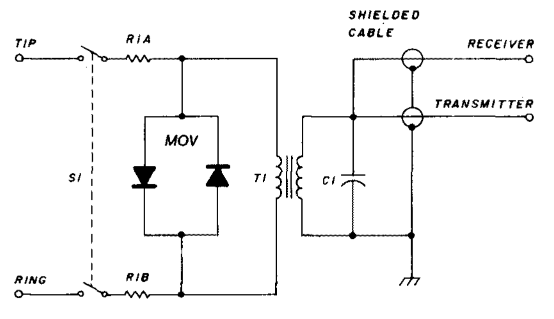

fig. 2. Basic phone patch.

The DC resistance of the transformer winding may be so low that it hogs most of the phone line current. Therefore, while using a phone in parallel for monitoring and dialing — which is recommended — the audio level on the incoming line may be too low. Resistors R1A and R1B (see **fig. 2**) will act as current limiters and allow the DC to flow through the phone where it's needed. If possible, these resistors should be carbon composition types.

To keep the line balanced, use two resistors of the same value and adjust the values by listening to the dial tone on a telephone handset. There should be little or no drop in volume when the patch transformer is switched across the phone line.

One of these transformers, or even two capacitors, can be used to patch two phone lines together, should there be a need to allow two distant parties to converse. There will be losses through the transformer so the audio level will degrade, but with two good connections this will not be a problem.

On the other side of the transformer — which could be called the secondary winding — choose one pin as the ground and attach the shields of the microphone and headphone cables to it. Attach the inner conductors to the other pin. The receiver output will work well into the 600-ohm winding, and if transmitting simplex or just putting receiver audio on the line there will be no crosstalk or feedback problems. In some cases, the audio amplifier in a receiver does not have enough output to feed the phone line at an adequate level; this can be handled by using the transformer with two secondaries (see the "improved" patch below) or by coupling a 8:1-kilohm transformer between the audio output and 600-ohm transformer. If RF is getting into the transmitter input, a capacitor (C1) across the secondary should help. A good value for the lower bands and AM broadcast interference is 0.1 μF. For higher frequencies, 0.01 μF usually gets rid of the problem. Unshielded transformers are sensitive to hum fields and building any patch into a steel box will help alleviate hum as well as RFI.

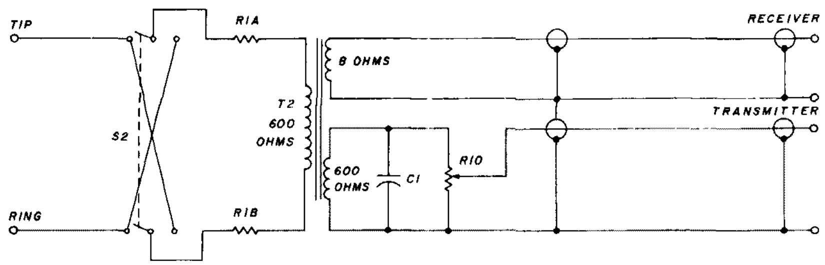

fig. 3. Improved phone patch.

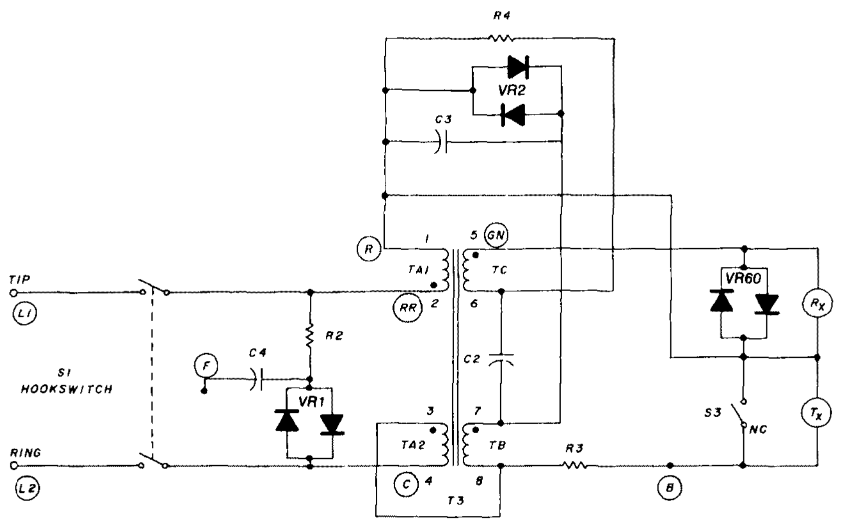

fig. 4. Typical network (United States). Note: circled letters are marked on network interconnection block terminals.

the improved phone patch

Several enhancements can be made to the basic phone patch to improve operation. The first is the addition of a double-pole double-throw switch to reverse the polarity of the phone line to reduce hum. This may not be necessary with a patch at the same location with the same equipment, but if it is, experiment with the polarity of the transformer connections and adjust for the least hum. Most of the time the balance will be so good that switching line polarity makes no difference. The switch should have a center "off" position or use a separate double-pole single-throw switch to disconnect from the line. The two secondaries on the "improved" patch (**fig. 3**) should be checked for balance by connecting the receiver and transmitter and checking for hum while transmitting and receiving. Switch the shield and inner conductors of the secondaries for minimum hum.

Many transmitters do not offer easy access to the microphone gain control. There may also be too much level from the patch to make adjustment of the transmit level easy. Placing R10 across the transformer allows easy adjustment of the level. It can be set so that when switching from the station microphone to the patch the transmitter microphone gain control does not need to be adjusted. This will also work on the basic 600-ohm 1:1 transformer. Most of the time a 1 kilohm potentiometer — logarithmic if possible — will work well. If not, a linear potentiometer will do. A 2.5-kilohm potentiometer may provide better control.

deluxe operation and VOX

Using VOX with a phone patch may cause a problem with receive audio going down the line and into the transmit input, triggering the VOX. There may not be enough Anti-VOX adjustment to compensate for this. The usual solution for this problem is to use a hybrid transformer, a special telephone transformer with a phasing network to null out the transmit audio and keep it off the receive line. Most telephones employ a similar transformer and circuit so that callers will not deafen themselves with their own voices. These devices are called "networks" (see **figs. 4** and **5**).

A network can be removed from an old phone and modified into a deluxe patch, or the phone can be left intact and connections made to the line and handset cords. The line cord should be coupled to a 600-ohm 1:1 transformer to keep the ground off the line. Note, in the network schematics, that the receiver and transmitter have a common connection; when coupling into radios or other unbalanced devices, make this the ground connection.

There may be confusion about terms used in the network. The telephone receiver is receiving the phone line audio, and the transmitter is transmitting the caller's voice. For phone patch use, a telephone receive line is coupled to the transmitter and the transmit line is coupled to the radio receiver. This is a fast way to put together a phone patch and may be adequate for VOX use.

A better patch can be built by using a network

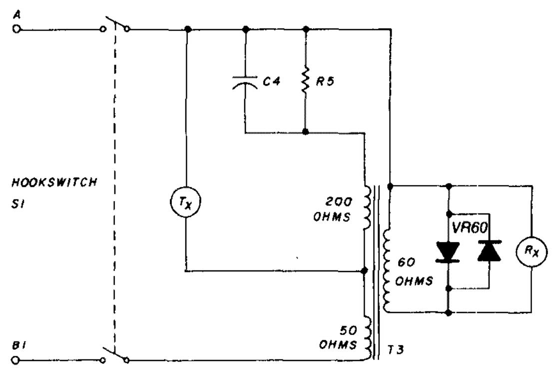

fig. 5. Typical European network.

removed from a phone or purchased from a local telephone supply house. This approach offers the added advantage of being able to adjust or null the sidetone. The circled letters in figs. 4 and 6 refer to the markings on the network terminal block. These letters are common to all United States networks made by Western Electric (AT&T), ITT, Automatic Electric, Comdial, Stromberg Carlson, and ATC.

To make the sidetone adjustable, remove R4 (R5 in European networks) and replace it with R11 (for European networks use R12). The Western Electric Network comes point-to-point wired and sealed in a can; the other networks are mounted on PCBs. To remove R4 from the Western Electric network, the can has to be opened by bending the holding tabs. Don't be surprised to find that the network has been potted in a very sticky, odious paste that has the texture of hot chewing gum and the odor of unwashed shirts. [This material — alleged to be manufactured according to a secret formula — will not wash off with soap and water. The phone company has a solvent for it, but because one of the secret ingredients is said to be beeswax, ordinary beeswax solvents such as gum turpentine, mineral turpentine (paint thinner or white spirit) and kerosene will work.] To remove the bulk of the potting compound, heat the opened can for 30 minutes in a 300 degree F (148 degree C) oven, or apply heat from a hot hairdryer or heatgun. You can also put the can out in the hot sun under a sheet of glass. Don't use too much heat because the plastic terminal strip may melt. Even with a film of compound remaining on it, the network can be worked on.

using a patch

For efficient use, a patch should have a telephone connected in parallel with it. This enables the operator to dial, answer, and monitor calls to and from the patch, as well as use the handset for joining in conversations or giving IDs.

One useful modification to the control telephone is adding a mute switch to the handset transmitter. This allows monitoring calls without letting room noise intrude on the line. It's also a good modification for high noise environments, where ambient noise enters through the handset transmitter and is heard in the receiver, masking the incoming call. Muting the transmitter makes calls surprisingly easy to hear. The mute switch can be a momentary switch used as a "Push-To-Talk" (PTT) or a Single Pole Single Throw (SPST) mounted on the body of the phone for long-term monitoring. The switch should be wired as Normally Closed, so that the transmitter element is muted by shorting across it (see fig. 4). This makes the mute "clickless." If the monitor phone uses an electret or dynamic transmitter it should still be wired as shown in fig. 4.

Transmit and receive levels on the phone line are a source of confusion that even telephone companies and regulatory agencies tend to be vague about. The levels, which can be measured in various ways, vary. But all phone companies and regulatory agencies aim for the same goals: enough level for intelligability, but not enough to cause crosstalk. The most trouble-free way to set the outgoing level on the patch is adjust the feed onto the phone line until it sounds slightly louder than the voice from the distant party on the phone line. If the level out from the patch is not high enough, the distant party will ask for repeats and tend to speak louder to compensate for a "bad line." In this case, adjust the level to the patch until the other party lowers his or her voice. The best way to get a feel for the level needed is to practice monitoring on the handset by feeding a broadcast station down the phone line to another Amateur who can give meaningful signal reports. It's difficult to send too much level down the phone while monitoring because the signal would simply be too loud to listen to comfortably. The major problem is sending too little signal down the line.

Coupling the phone line into the radio transmitter is not much more difficult than adjusting a microphone to work with a radio transmitter. Depending on the setup, the RF output indication on a wattmeter, the ALC on the transmitter or even listening to the transmitted signal on a monitor receiver will help in adjusting the audio into the radio transmitter. Phone lines can be noisy, and running too much level into the transmitter and relying on the ALC to set the modulation can cause a fair amount of white noise to be transmitted. Watching the RF output while there are no voice or control signals on the line will help in adjusting for this. VOX operation can alleviate the problem of noise being transmitted during speech pauses.

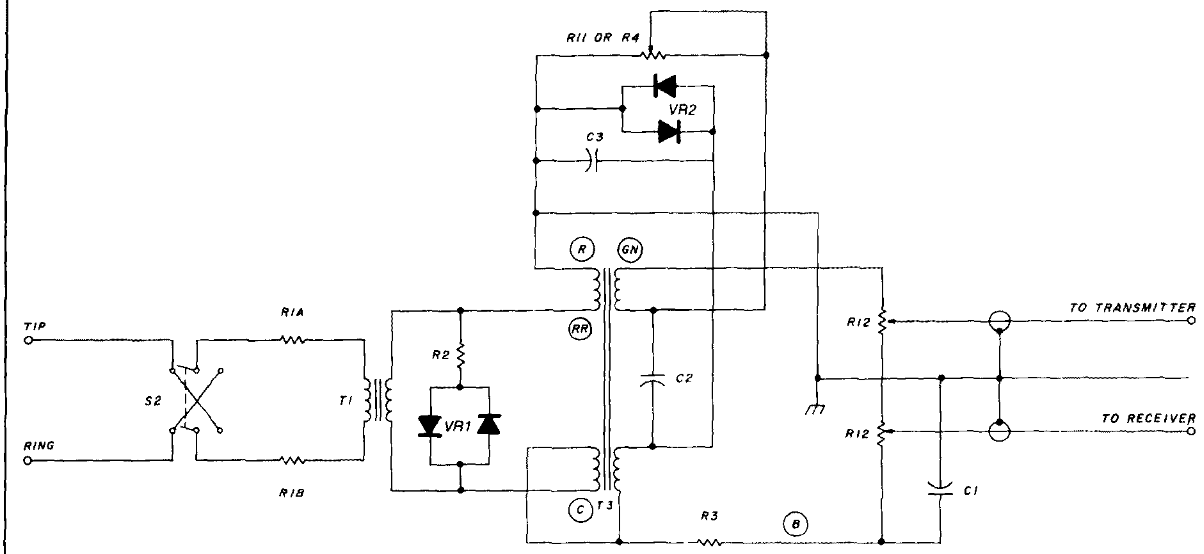

fig. 6. Deluxe phone patch. (Note: while *ham radio* designates varistors as *CR*, telephone companies customarily use the *designation VR*. — Ed.)

item	description
C1	0.1 μF (see text)
C2	1.5 to 2.0 μF (depending on manufacturer)
C3	0.47 μF not used in all networks
C4	0.1 μF
C5	2.0 μF 250 volt mylar film (see text)
MOV	130 to 250 volt MOV (see text)
R1A,B	100 to 270 ohms (see text)
R2	180 to 220 ohms (depending on manufacturer)
R3	22 ohms
R4	47 to 110 ohms (depending on manufacturer)
R5	1 kilohm
R6	1 kilohm (see text)
R7	10 ohm (see text)
R10	1 kilohm potentiometer (see text)
R11	200 ohm potentiometer (see text)
R12	2 kilohm potentiometer (see text)
S1	DPST or hookswitch
S2	DPDT, center off (see text)
S3	NC momentary switch (see text)
T1	600 ohm 1:1 transformer
T2	600 ohm primary, 600 ohm and 8 ohm secondary (see text)
T3	network transformer
VR1,VR2 VR60	silicon carbide varistor or back-to-back zener

fig. 6. Parts list.

A hybrid patch used for VOX operation needs to be adjusted carefully for good performance. If it has a null adjustment, this should be set before adjusting the VOX controls. Using a separate receiver/transmitter setup is the easiest way to adjust the patch. The phone line should be attached to a silent termination: the easiest way to do this is to *dial part of a number*; another way to do it is call a cooperative friend. Tune the shack receiver to a "talk" broadcast station or use the BFO as a heterodyne. With the transmitter keyed into a dummy load, set the null adjustment potentiometer R11 (R12 for European phones) for a minimum RF output on the transmitter. Using a transceiver, place an oscilloscope or audio voltmeter across the microphone input terminals and, while receiving a signal, adjust for the lowest voltage. For proper operation, it's important that the phone be connected to the patch during these adjustments since the hybrid relies on all inputs and outputs being terminated.

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ham radio

passive audio filter design, part 2: highpass and bandpass filters

A highpass filter with high attenuation at 60 and 120 Hz — and some common bandpass filter design problems

The simplest highpass filter possible is a single capacitor. But if you think that a capacitor is fairly useless as a highpass filter, consider the circuit shown in **fig. 1**. A voltage source, V_S, of resistance R_S, is driving a low-pass filter through a capacitor, C_H. The source and termination impedances of the filter are R_S and R_T respectively, which we will assume are equal.

As the frequency (of the voltage source) increases, the reactance of C_H decreases. Eventually a point is reached where the reactance of C_H is so low that it is insignificant compared to R_S. The combination of C_H and the low-pass filter will therefore have a bandpass response in which C_H attenuates low frequencies and the low-pass filter attenuates high frequencies. The low frequency attenuation will, of course, be fairly modest but can have a useful effect when applied to a filter such as the practical 1-dB/50-dB elliptic low-pass described earlier.[1]

Figure 2 shows simulated results of how the response of the 1-dB/50-dB filter is modified by various preferred values of C_H. Also shown for comparison is the unmodified response; that is, with C_H short-circuited. With $C_H = 0.22\ \mu F$, considerable attenuation of low frequencies is obtained, but C_H is still effective (i.e., it introduces a reactive term) at almost 2 kHz, which considerably narrows the bandwidth of the total network. As C_H increases, the response above 2 kHz follows the low-pass response more closely, and the low-frequency attenuation is reduced. One useful value of C_H is 2.2 μF, which produces a significant reduction in the passband ripple when compared with the unmodified low-pass response. The mismatch in

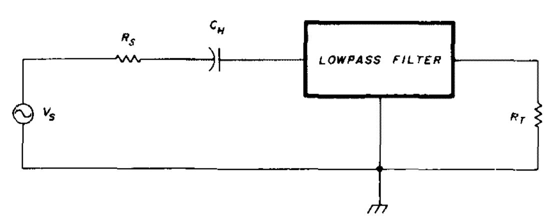

fig. 1. Combining a series capacitor and low-pass filter provides composite bandpass response.

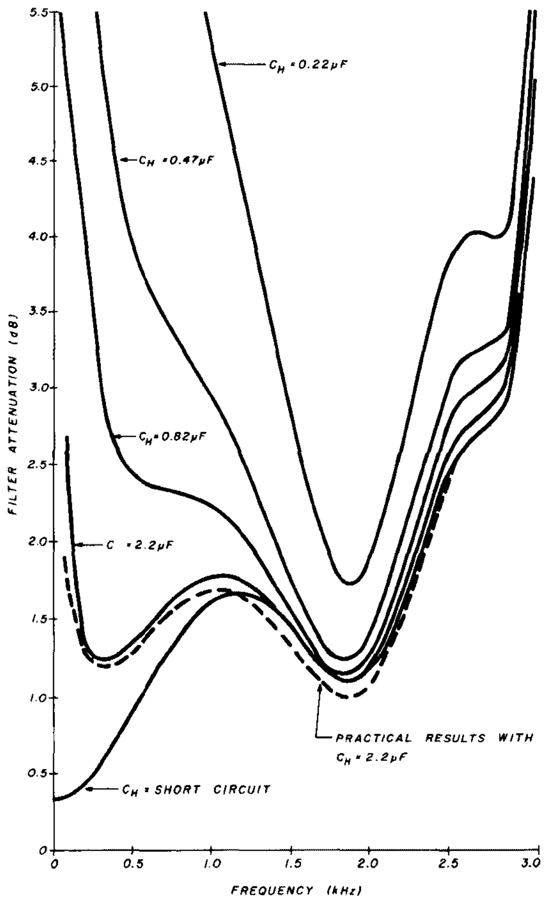

fig. 2. 0-3 kHz response of 1-dB/50-dB five-branch elliptic low-pass filter with various values of input capacitor, C_H showing simulation and practical results.

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the drive impedance to the filter caused by the reactance of C_H does not cause a drastic degradation in performance. Only the frequency-dependent variable effect by C_H is evident in the combined response.

Practical results with C_H equal to 2.2 μF are also plotted in **fig. 2**, and they show close agreement with the simulated results. Although no great attenuation of unwanted low frequencies (such as 60 Hz and 120 Hz) is obtained, it is recommended that the 1-dB/50-dB low-pass filter be used in conjunction with a 2.2 μF input capacitor to reduce the passband ripple. A compact Siemens metalized polyester 2.2 μF capacitor* is available and is preferable to a polarized capacitor, which would have a greater tolerance.

I will now describe an improved highpass filter that provides high attenuation at 60 Hz and 120 Hz and at unwanted lower speech frequencies. **Figure 3** shows the circuit diagram of the filter, with and without resistors to simulate the low-Q inductors.

This filter is a five-branch Butterworth highpass with a theoretical response rolloff of 30 dB per octave. Highpass filters are generally designed by transforming a lowpass prototype, and this procedure is described in appendix A of this article.

Table 1 shows the component values of the original 1-ohm, 1-rad/sec low-pass filter;[2] the 1-ohm, 1-rad/sec transformed highpass; the scaled 500-ohm, 500-Hz highpass with theoretical values and ideal inductors; the 500-ohm, 500-Hz highpass with rounded values

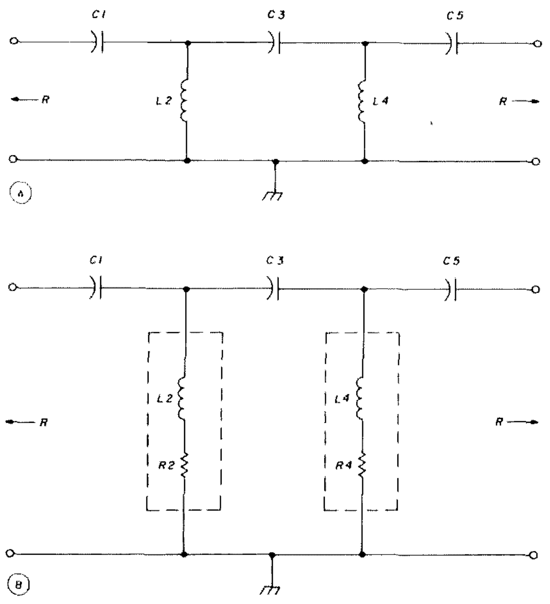
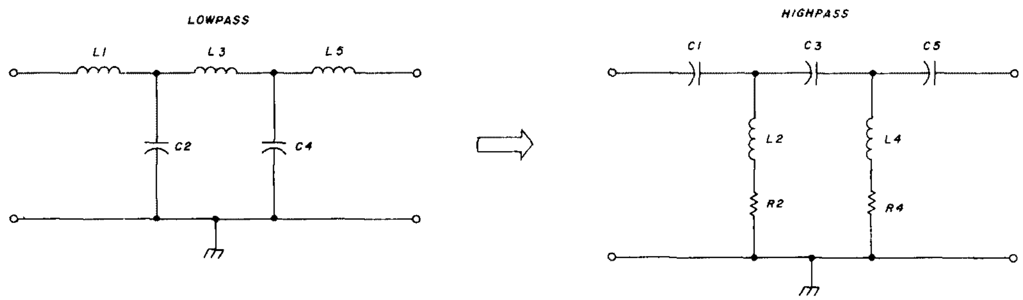


fig. 3. Butterworth five-branch highpass filter. (*A*) Schematic of filter with ideal inductors, (*B*) schematic of filter with real inductors.

*Digi-Key part no. E1225, 2.2 μF/100V.

table 1. Component values of Butterworth five-branch highpass filter. 500 ohm, 500-Hz values are obtained by multiplying 1 ohm, 1 rad/sec values by 6.366×10^{-7} for capacitors and 0.1592 for inductors.

low-pass prototype component	1 ohm, 1 rad/sec value	highpass transformed component	1 ohm, 1 rad/sec value	500 ohm, 500 Hz theoretical value ideal inductors	500 ohm, 500 Hz rounded value ideal inductors	500 ohm, 500 Hz rounded value real inductors
L1	0.618 H	C1	1.618 F	1.030 μF	1 μF	1 μF
C2	1.618 F	L2	0.618 H	98.34 mH	100 mH	100 mH
L3	2.000 H	C3	0.500 F	0.3183 μF	0.33 μF	0.33 μF
C4	1.618 F	L4	0.618 H	98.34 mH	100 mH	100 mH
L5	0.618 H	C5	1.618 H	1.030 μF	1 μF	1 μF
		R2	—	0 ohms	0 ohms	82 ohms
		R4	—	0 ohms	0 ohms	82 ohms



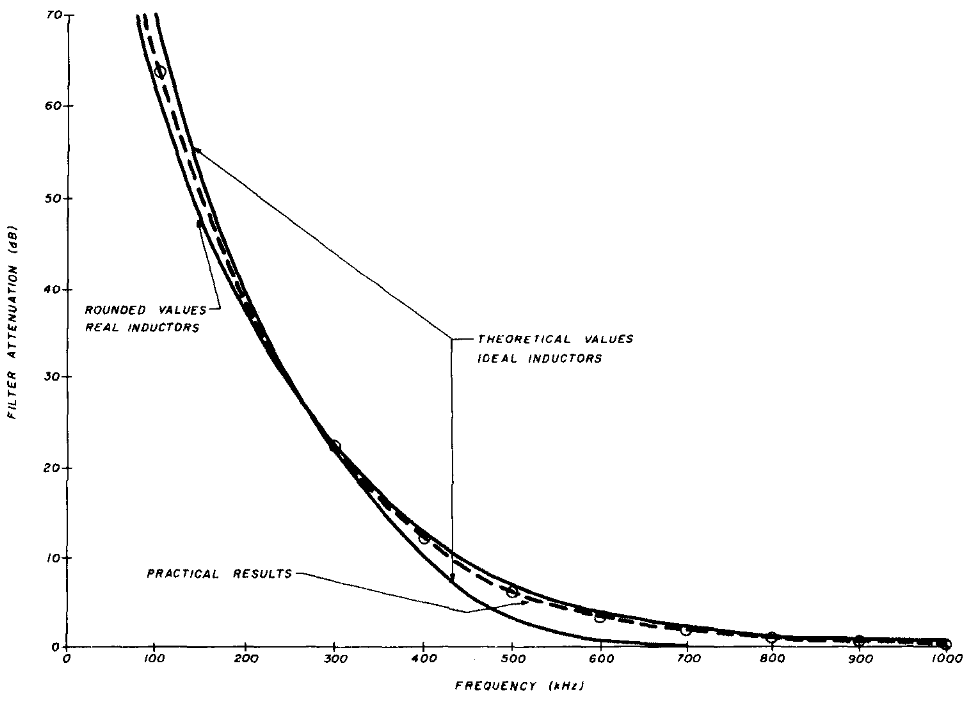


fig. 4. 0-1 kHz response of 500-Hz Butterworth highpass filter showing simulation and practical results. (*A*) theoretical values, ideal inductors; and (*C*) practical results —O— —O— —O—.

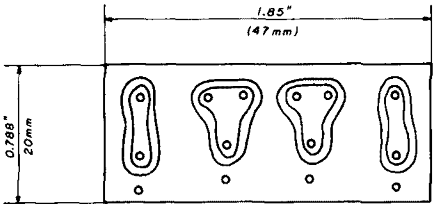
and ideal inductors; and the 500-ohm, 500-Hz highpass with rounded values and real inductors.

A cutoff frequency (defined as the 3-dB attenuation frequency for Butterworth filters) of 500 Hz means that any 120-Hz input component (which is more than 2 octaves below the cutoff frequency) will be attenuated by more than 60 dB. Any 60-Hz input will be attenuated (theoretically) by more than 90 dB, as it is another octave below 120 Hz. More than half the total power of speech lies below 450 Hz,[3] so using a 500 Hz cutoff highpass filter at the audio input to a transmitter will result in a considerable saving in power. Intelligibility of speech will not be influenced; however, for better quality speech, a cutoff frequency of perhaps 300 Hz would be more appropriate. If desired, the reader can scale the 1-ohm, 1-rad/sec highpass values to 500 ohms, 300 Hz, by multiplying the capacitors by 1.061×10^{-6} and the inductors by 2.65×10^{-1}. Rounding the answers will then give practical values for the components.

The 0 to 1 kHz response of the 500-ohm, 500-Hz highpass filter is shown in **fig. 4**. Curve A is the theoretical value, real-inductor response and the 3-dB attenuation frequency is 500 Hz, as predicted. The 30 dB and 60dB attenuation frequencies are 250 Hz and 125 Hz respectively, giving the classical five-branch Butterworth response.

A curve of the rounded values, ideal-inductor response has not been plotted in **fig. 4** because simulations indicated that it deviated from curve A by no more than 0.8 dB at any frequency.

Curves B and C are the simulated rounded values, real-inductor response, and the measured response, respectively. These two curves follow each other closely. One effect of the low Q inductors is rounding



Components for 500-ohm, 500-Hz Butterworth highpass filter:
C1 1 μF Siemens metelized polyester 5 percent
C3 0.33 μF Siemens metelized polyester 5 percent
C5 1 μF Siemens metelized polyester 5 percent
L2 100 mH Toko 10RB
L4 100 mH Toko 10RB

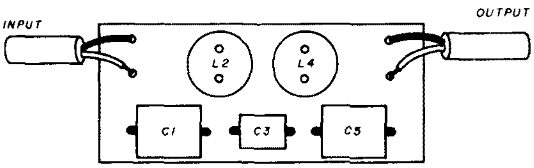


fig. 5. PC board foil pattern and component placement for five-branch Butterworth highpass filter.

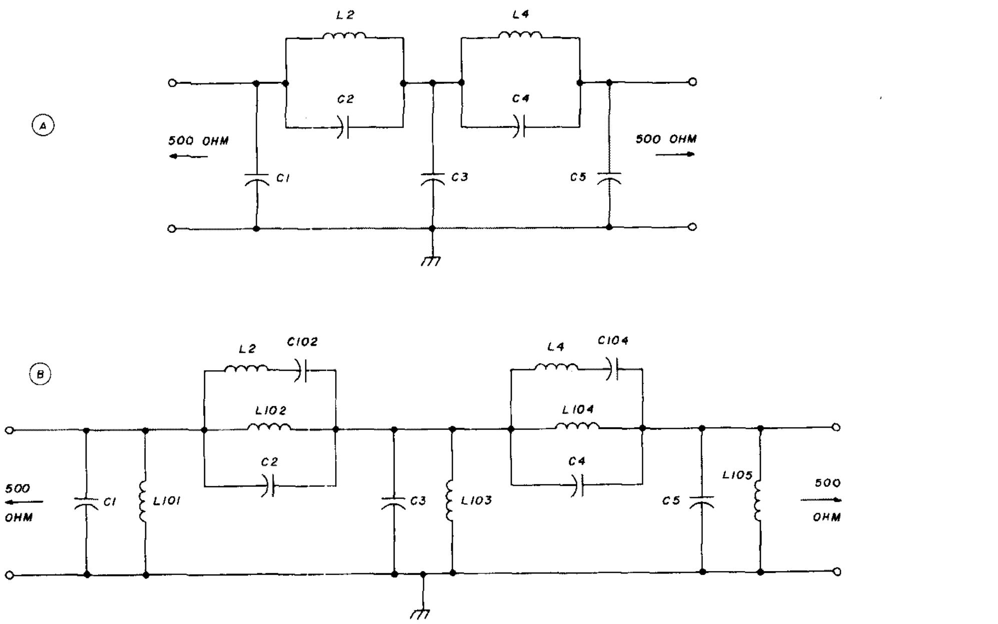


fig. 6. Elliptic bandpass filter transformation procedure. (*A*) Schematic of 1-dB/50-dB 3-kHz lowpass prototype, (*B*) schematic of 1-dB/50-dB 300-3300 Hz bandpass filter.

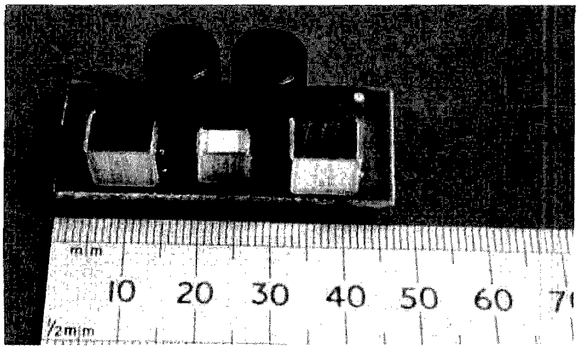


fig. 7. Experimental five-branch Butterworth highpass filter.

of the passband edge, the attenuation at 500 Hz being 6 dB rather than the theoretical 3 dB. The filter also still has a loss of approximately 1 dB at 1 kHz.

Figure 5 shows the PC board foil pattern and component layout for this highpass filter, and **fig. 7** is a photograph of my prototype.

bandpass filters

The modern approach to bandpass filter design is by transformation of a low-pass prototype, as described in appendix B.[4] This method results in a symmetrical response and if the prototype low-pass filter were elliptic, then the bandpass will also be elliptic, having passband ripple and minima of attenuation in the lower and upper stopbands.

I want to demonstrate some drawbacks of this technique, using the 1-dB/50-dB lowpass filter described earlier as the prototype. The specification of the bandpass filter to be designed is:

passband ripple (A_P)	1dB
stopband minimum attenuation (A_S)	50 dB
ripple cutoff frequencies	300 Hz, 3300 Hz
50-dB stopband attenuation bandwidth	4221 Hz
source impedance	500 ohms
load impedance	500 ohms

The following points should be noted about this filter. First, the passband bandwidth of the bandpass filter (3 kHz) is the same as that of the low-pass prototype. Secondly, the 50-dB stopband attenuation bandwidth of the passband filter is the same as that of the low-pass prototype (4221 Hz). Thirdly, the passband ripple (A_P) and minimum stopband attenuation (A_S) are identical to those of the prototype. Normally, for speech processing, an upper cutoff frequency of 3 kHz would be chosen, but I have not selected a 2700-Hz low-pass prototype for transformation because I want to illustrate only the problems that can be encountered with this approach, not produce a practical design.

Figure 6 shows the schematics of the low-pass prototype and the final, transformed bandpass filter.

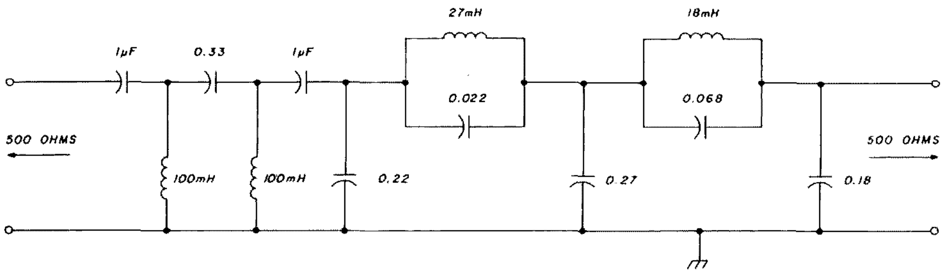


fig. 8. Schematic of the bandpass filter formed from the 500-ohm, 500-Hz Butterworth and 1-dB/50-dB elliptic lowpass filters.

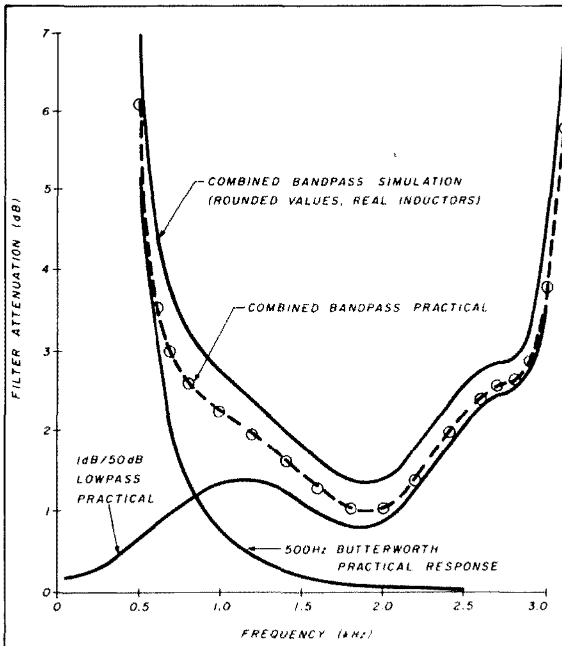


fig. 9. 0-3 kHz response of 500-Hz Butterworth highpass filter with 1-dB/50-dB five-branch elliptic low-pass filter, showing simulation and practical results. (Low-pass and highpass responses also shown.)

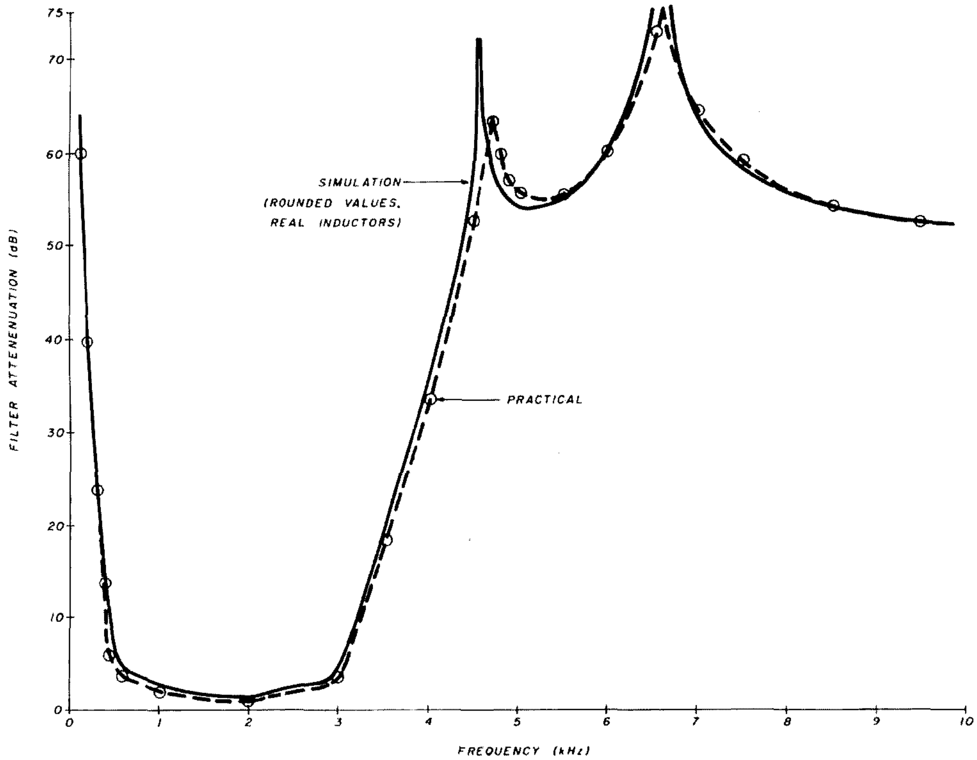


fig. 10. 0-10 kHz response of 500-Hz Butterworth highpass filter with 1-dB/50-dB five-branch elliptic low-pass filter, showing simulation.

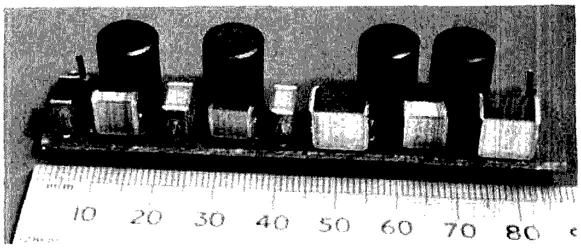


fig. 11. Experimental 500-Hz Butterworth highpass filter with 1-dB/50-dB five-branch elliptic low-pass filter.

Capacitors C1 through C5 have inductors L101 through L105 in parallel and inductors L2 and L4 have capacitors C102 and C104 connected in series. The value of each additional component has been calculated to resonate with the original component at the geometric mean of the lower and upper ripple cutoff frequencies. That is, if the lower and upper ripple cutoff frequencies are $f1$ and $f2$, then the geometric mean, $f0$ is given by:

$$f0 = \sqrt{f1 \cdot f2}$$

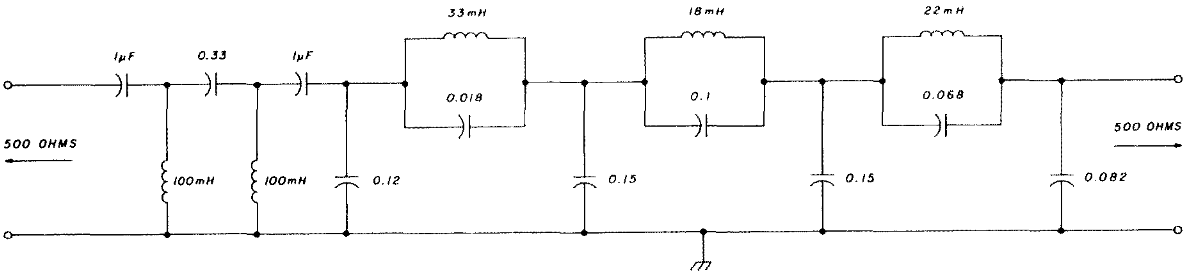


fig. 12. Schematic of the bandpass filter formed from the 500-ohm, 500-Hz Butterworth and 0.18-dB/50.1-dB elliptic low-pass filters.

For this filter, *f0* is equal to 995 Hz. Taking C1 (0.2051 μF) as an example, L101 is made equal to 124.7 mH, giving the resonant frequency of 995 Hz. **Table** 2 shows the low-pass prototype and bandpass filter component values.

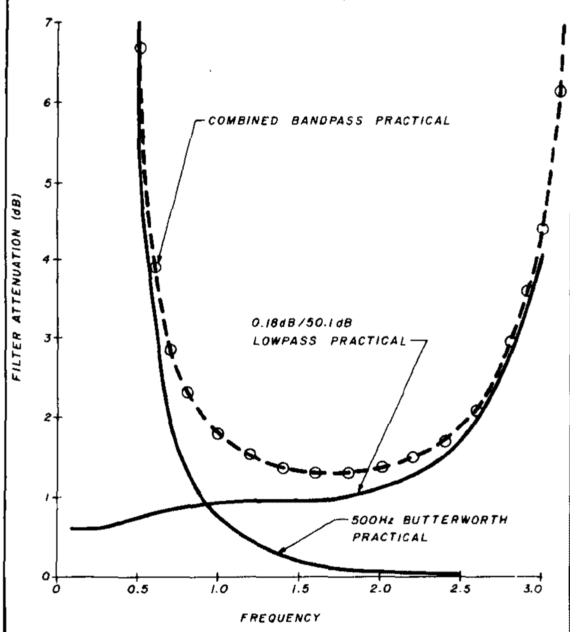


fig. 13. 0-3 kHz response of 500-Hz Butterworth highpass filter with 0.18-dB/50.1-dB seven-branch elliptic low-pass filter, showing measured response. (Low-pass and high-pass responses also shown for comparison.)

The simulated bandpass filter has the expected response, indicating a valid design procedure. There are methods of simplifying the circuit by further transformation, but these considerably complicate the design procedure.

There are two problems with this bandpass filter design technique, however. The first is that it results in several awkward value components. In this example, L102 (1.08 H) and L104 (385.2 mH) would be difficult to wind if homemade inductors were used. C102 (1.003 μF) and C104 (1.28 μF) are also rather high values. This problem becomes worse if a more complex low-pass prototype is used. Consider, for example, the 0.18-dB/81-dB low-pass filter previously described: choosing an inductor to resonate at 995 Hz with C2 (6897 pF) would require a value of 3.71 HI

The second problem is caused by the filter's symmetry of response. In the example considered, the low-frequency response is probably too good for speech filtering applications, and the penalty paid is the excessive number of components (seven capacitors and seven inductors) required for the final filter. It would be useful if the high- and low-frequency responses could be selected separately, both in rolloff rate and type. For example, for speech filtering, and rapid rolloff elliptic low-pass response would be ideal, along with a more modest Butterworth highpass response. This is exactly my approach to bandpass filter design. It is generally thought that intermediate buffering

table 2. Component values of 1-dB/50-dB elliptic low-pass prototype and transformed bandpass filter.

3 kHz low-pass prototype		300-3300 Hz bandpass filter	
component	**value**	**component**	**value**
C1	0.2051 μF	C1	0.2051 μF
		L101	124.7 mH
C2	0.0237 μF	C2	0.0237 μF
		L102	1.081 H
C3	0.2538 μF	C3	0.2538 μF
		L103	100.8 mH
C4	0.0664 μF	C4	0.0664 μH
		L104	385.2 mH
C5	0.1735 μF	C5	0.1735 μH
		L105	147.2 mH
L2	25.5 mH	L2	25.5 mH
		C102	1.003 μF
L4	19.9 mH	L4	19.9 mH
		C104	1.286 μF

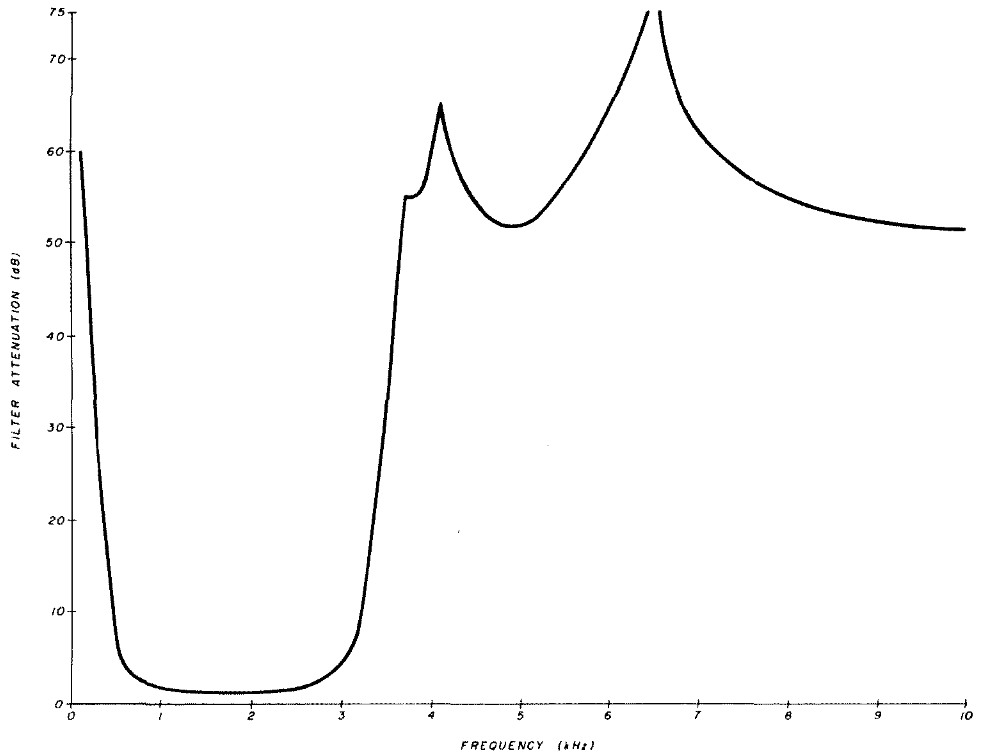


fig. 14. 0-10 kHz response of 500-Hz Butterworth highpass filter with 0.18-dB/50.1-dB seven-branch elliptic low-pass filter, showing practical results only.

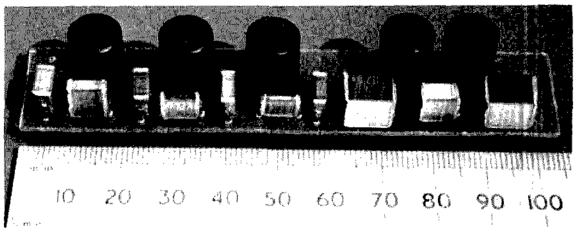


fig. 15. Experimental 500-Hz Butterworth highpass filter with 0.18-dB/50.1-dB seven-branch elliptic low-pass filter.

should be used between two filter types, but I will demonstrate that this is unnecessary and that compact bandpass filters can easily be designed and constructed.

Since three low-pass and one highpass (excluding the single capacitor) filters have already been described, they will be used to form bandpass filters. **Figure 8** shows the schematic of the 500-ohm, 500-Hz Butterworth highpass cascade with the 1-dB/50-dB elliptic low-pass filter. The component values shown are rounded from the original filters.

Figure 9 shows the 0 to 3-kHz response and **fig. 10** the 0 to 10-kHz response of the resulting bandpass filter. Only the simulated response with rounded values and real inductors is plotted, along with the practical results obtained. The practical results for the highpass and low-pass sections are also shown in the range 0 to 3 kHz for comparison with the bandpass response. Below approximately 750 Hz, the bandpass filter follows predominantly the highpass section response, and above 1.5 kHz the response is predominantly that of the low-pass section. Between these frequencies, the response is very nearly the algebraic sum of the separate sections. The practical bandpass filter has 3-dB cutoff points (measured with respect to the insertion loss of 1 dB at 1.9 kHz) of approximately 675 Hz and 3 kHz.

A detailed PC board foil pattern and component layout are not shown for this filter, but a photograph of the completed filter is shown in **fig. 11**. I simply took the two original layouts (**part 1, fig. 7** and **part 2, fig. 5**) and joined them together. Incidentally, it makes no difference to the response whether the highpass or low-pass section comes first.

The same procedure has been applied to the 500-ohm, 500-Hz Butterworth highpass and the 0.18-dB/50.1-dB elliptic low-pass filters in the schematic shown in **fig. 12**. The response of the resulting bandpass design is shown in **figs. 13** and **14**. Reduced passband

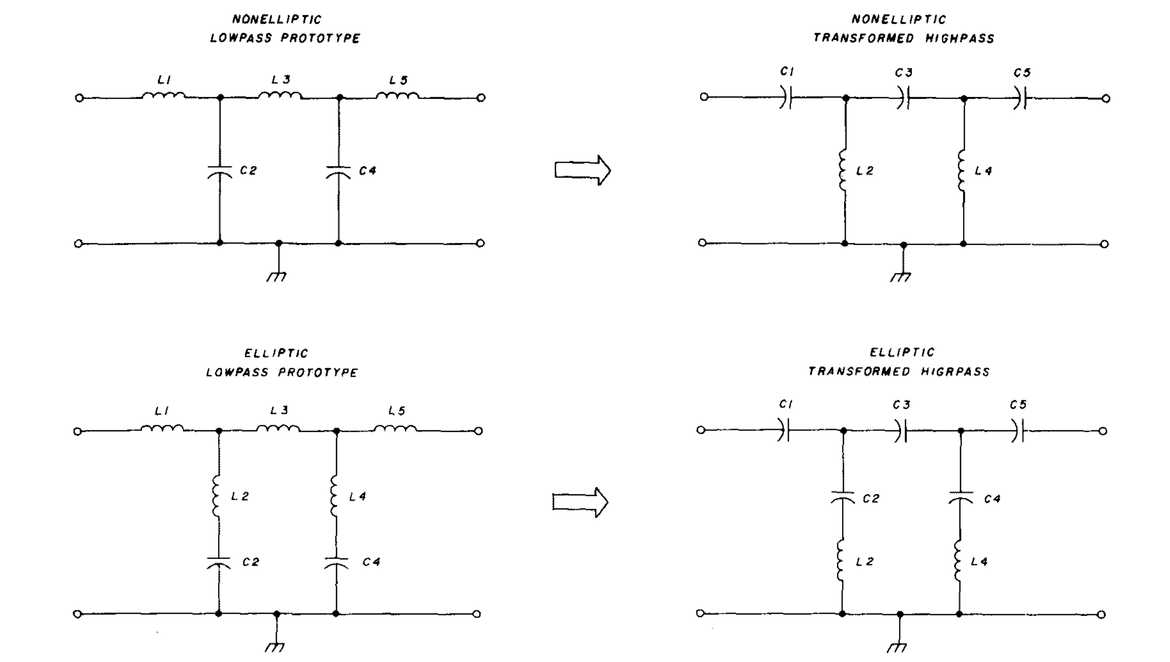


fig. A1. Schematics of low-pass-to-highpass transformations showing (A) non-elliptic filter, and (B) elliptic filter.

ripple and improved low-pass rolloff are evident because of the superior low-pass filter used.

A photograph of this bandpass filter is shown in **fig. 15**. Again the PC board layout is a combination of the two separate filter layouts (**part 1, fig. 13**, and **part 2, fig. 5**.) Either of these two bandpass filters would form an excellent post-detector filter in a superhet or direct conversion receiver.

appendix A
low-pass to highpass transformation

Figure A1 shows a typical nonelliptic low-pass-to-highpass filter transformation and a typical elliptic low-pass-to-highpass filter transformation. Note that in both cases the minimum capacitor implementation has been chosen for the low-pass prototype, so that minimum inductor design will result for the highpass filter.

In both cases, the new component values are obtained by use of the formulas:

$$C1\ (highpass) = \frac{1}{L1\ (low\text{-}pass)}$$

$$L2\ (highpass) = \frac{1}{C2\ (low\text{-}pass)}$$

where capacitors and inductors have their 1 ohm, 1-rad/sec values. The final values are obtained by scaling to the desired highpass filter impedance and cutoff frequency, as shown in part 1, appendix A.

appendix B
low-pass-to-bandpass transformation

A low-pass prototype is first selected which has the same bandwidth as the desired bandpass response. The geometric mean ($f0$) of the lower ($f1$) and upper ($f2$) cutoff frequencies for the bandpass filter is then determined by:

$$f0 = \sqrt{f1 \cdot f2}$$

Each capacitor in the low-pass prototype is then transformed into a capacitor/inductor parallel combination as shown in **fig. B1A** and each inductor is transformed into an inductor/capacitor series combination, as in **fig. B1B**. In the case of the capacitor transformation, the new inductor value is chosen to resonate with the original capacitor at $f0$. In the case of the inductor transformation, the new capacitor value is chosen to resonate with the original inductor at $f0$.

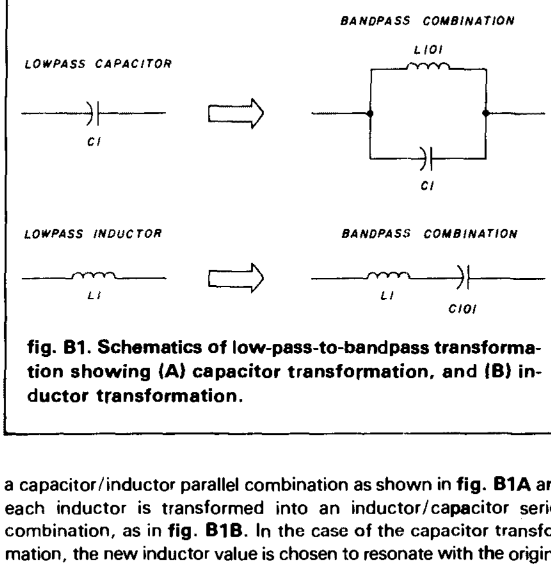


fig. B1. Schematics of low-pass-to-bandpass transformation showing (A) capacitor transformation, and (B) inductor transformation.

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ham radio

build a
fail-safe digital clock

Don't lose count when the power fails

Although digital clocks have simplified keeping track of time accurately, they present some problems not encountered with synchronous-motor analog clocks. When line power is lost, for example, the mechanical display of motor-driven clocks provides a non-volatile memory that stores the time at the instant of power loss. Once line power returns, the motor restarts and continues counting the cycles. Unless an extremely accurate measurement of time is needed (for keeping a meteor scatter schedule, for *example*) the clock must be reset only when major power failures lasting more than a few minutes occur. With digital clocks that run on line power, however, even a momentary power dropout results in a complete loss of time.

alternatives for standby power

With synchronous-motor-driven analog clocks there's no simple way of providing for a standby power source when line power is lost. Electronic clocks, on the other hand, offer alternatives. You can, of course, power the clock with a battery, but this creates other problems: over long periods the time-base oscillator frequency error will cause any error in time to increase. Crystal aging, frequency sensitivity to changes in battery voltage and temperature, as well as the initial accuracy of the oscillator frequency all contribute to timing errors that will eventually require the clock to be reset.

A better choice is to combine the best features of line power and battery operation. Line power can be used to provide an accurate long-term time base, keep the battery charged, and allow the use of a light-emitting diode (LED) display. Battery power can be used to power the clock when line power is lost, with a crystal-oscillator time base for keeping time. A logic circuit detects the presence or absence of line power and puts the clock in the appropriate operating mode. Additional logic allows you to override the automatic changing of the clock time base and run with the crystal oscillator for accurate short-term time keeping. The clock design presented here has these features and has proven itself in over a year of continuous operation.

contribution options

Besides building the entire clock — including the crystal oscillator time base, logic, circuit and battery backup — as a project, the design information can be used for any clock using the MM5314 IC or its close relatives, the MM5309, MM5311, MM5512, MM5313, and MM5515. Existing clocks using this IC can be modified to add the crystal oscillator, logic, circuit, and battery without great difficulty. Another option is to start with a digital clock kit and simply add the new components*.

Others may want the clock to run only on the crystal-oscillator time base rather than switch between it and line power. The numerous design alternatives available allow you to choose the features that you want.

clock functions

The different functions of the clock are controlled by the logic circuit (see **table 1**). A schematic is shown in **fig. 1**. The crystal oscillator and divider IC run con-

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*Two kits that provide a starting point are No. JE701 and No. JE747, from Jameco Electronics, 1355 Shoraway Road, Belmont, California 94002.

table 1. Logic control definitions.

function	control	logic 1	logic 0
run/hold	S1	run	hold
slow set	S2	normal	slow set
fast set	S3	normal	fast set
4/6 digit	S4	4 digits	6 digits
time base	S5A	normal	display enable
	S5B	normal	crystal oscillator
display enable	S6	normal	enabled
12 hr/24 hr	U1-pin 10	24 hour	12 hour
50 Hz/60 Hz	U1-pin 11	50 Hz	60 Hz

tinuously to eliminate any startup and shutdown transients that would degrade time-keeping accuracy. The missing pulse detector monostable selects either line power or the oscillator 60-Hz signal for the clock time base, based on the presence or absence of line power. A switch input to the logic allows the monostable control to be overridden so that the crystal oscillator provides the clock time base even when line power is present.

To conserve battery power the logic normally disables the display when line power is lost. A nonlatching, normally open pushbutton switch enables the display during battery operation so that the time can be read if desired. A two-pole switch selects the crystal oscillator time base. One pole disables the 60-Hz signal from line power so that the monostable times out and allows the 60-Hz signal from the oscillator/divider to run the clock. The other pole parallels the display enable switch and keeps the display on when running in the crystal-oscillator time-base mode.

The run/hold switch stops the clock from counting when setting the time. Two nonlatching, normally open pushbutton switches allow fast and slow time set: fast set advances the time at one hour per second; slow set advances it at one minute per second. Another switch selects either a four-digit (hours and minutes) or a six-digit (hours, minutes, and seconds) time display. The seconds display is necessary when setting the clock, but is distracting when logging or recording time; more than once I've found myself writing down the wrong four digits on the log sheet.

The clock IC can display time in either a 24-hour or 12-hour format. If a 12-hour format is desired, pin 10 of U1 should be grounded. Two LEDs are used as colons between the hours digit and tens of minutes digit on the display. Besides giving the unit a more clock-like appearance, the LEDs serve as pilot lights. When line power is lost, only the display digits are disabled by the clock IC. The colons remain lit to show that the clock is in the battery mode and operating.

circuit details

Power supply. The AC line circuit follows standard design and safety practices. A three-wire plug and cord are used, with the ground wire (green) connected to the clock's metal case. A transient suppressor, V1, protects the MOS and CMOS ICs from high amplitude voltage transients. C1 filters transient high-frequency components as well as any RF that may be on the power wiring. (If your line power is extremely contaminated with transients, it may be necessary to add small inductors to the power lines to give the capacitor a higher impedance to work against.) I included a fuse to prevent failure in the clock from creating a fire hazard and protect the transient suppressor from failing under long-term high-voltage conditions.

The transformer has a 12.6 volt RMS secondary with a current rating of 200 mA. The rectified output voltage under load should be in the range of 11 to 16 volts. The LED display draws the major portion of the current in the clock. The 200 mA rating is the minimum for the transformer because of LED display loading. Higher current transformers should be used if space is available. A diode bridge rectifier is used with the filtering provided by C2. The filter capacitor also makes the transition between line power and battery power smoother by slowing down the voltage rate of change. A smaller value capacitor, C3, is in parallel with the filter capacitor to bypass any residual power transient high-frequency components or RF signals. Another capacitor, C17, provides additional bypassing.

Battery supply. The NiCad battery circuit also includes a trickle charging resistor, R1, and a blocking diode, CR5. R1 gives a trickle current equal to a hundredth of the battery capacity. The battery has a capacity of 65 mA per hour, which requires a trickle current of 0.65 mA to maintain full battery charge. The normal slow charge rate is 7mA for 14 hours. If other types of rechargeable batteries are used, be sure to change the value of R1 to obtain the correct trickle charge current.

When the battery is powering the clock, the current will flow through CR5 to avoid the high resistance path through R1. A diode keeps the voltage drop to a minimum. With the time display disabled, the clock circuit draws about 8 mA, so approximately 8 hours of continuous battery operation is possible. The absolute minimum supply voltage for the clock is probably under 5 volts, so that the readily available 9 volt NiCad battery can be used.

Clock integrated circuits. The clock IC requires external components to drive and multiplex the LEDs in the time display. This multiplexing is necessary to reduce the peak current drawn by the display and reduce the number of pins required on the IC. If the six digits with their seven segments were driven direct-

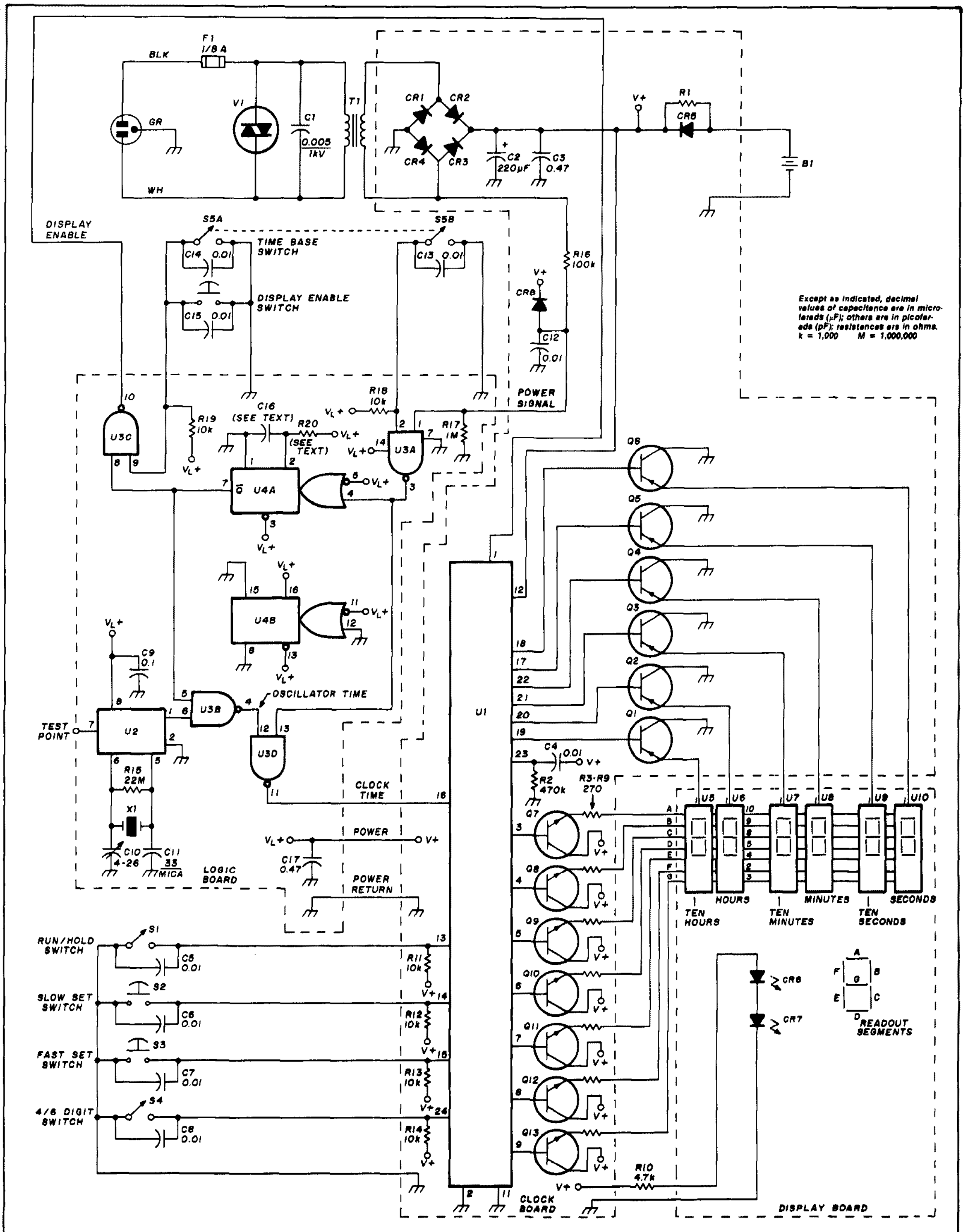

fig. 1. Schematic diagram of the digital clock, with battery backup and crystal oscillator time base.

ly, 42 logic lines would be required. By wiring all the digit segments in parallel and multiplexing, only 13 logic lines are needed. Seven outputs are for the seven segments in each LED digit and six outputs for the six digits. The logic inside the clock IC keeps track of which digit is being enabled and turns on the correct segments. The digits are scanned at a rate of approximately 1 kHz. The multiplex frequency is determined by C4 and R2, and is approximately equal to 3/(R2xC4). The display driver devices are general-purpose switching transistors. The LED segment currents are determined by R3 through R9.

The external clock controls have bypass capacitors across their switches to reduce any RF pickup or switch bounce transients. The pullup resistors at the clock IC allow the bypass capacitors to follow the operating voltage when changing between line supply and the battery. The input resistances of the clock IC are high because of the P channel MOS construction.

The capacitor holds the lower battery voltage while the clock IC is running on the higher line supply voltage, and misinterprets that as a logic 0 instead of a logic 1. Before the pullup resistors were added, the clock gained exactly one minute every time it went from battery to line supply power because of the capacitor storage time. Since the capacitor memory time depends on its value and the input resistance of the clock IC, it was easier to add the pullups than to fine tune each clock for proper operation. The pullup resistor values are much smaller than the clock IC input resistance. The time constant is 0.1 millisecond to prevent false logic conditions caused by power transitions.

Oscillator/divider. The oscillator/divider, U2, provides a crystal oscillator-based 60-Hz output. The division ratio is fixed at 59,659, so a variable capacitor sets the oscillator frequency to the exact value of 3,579.540 kHz. A buffered oscillator output is available when setting the frequency to avoid test equipment loading effects. If you don't have an accurate frequency counter to set the oscillator, or if crystal time-base accuracy is not critical, variable capacitor C10 can be replaced with a fixed 33 pF mica capacitor. To gauge the accuracy you need, remember that a 1-Hz oscillator frequency error will amount to a 8.81-second time error over the course of a year. Crystal frequency tolerances will be about ±300 Hz from their nominal value, with aging contributing a long-term downward drift of 3-7 Hz/year. The frequency of the oscillator also depends on its supply voltage. My unit had a sensitivity of approximately 10 Hz/volt over the range of 6 to 12 volts. If battery backup is your main operating mode, adjust the crystal oscillator frequency while running on battery power.

Logic. The 60-Hz line power time-base signal is taken from one side of the power transformer secondary. A low-pass filter consisting of R16 and C12 attenuates high-frequency transient components that might get into the clock circuit and the time base. A diode clamp, CR8, protects the input to U3 by preventing the input voltage from exceeding the supply voltage by more than a diode drop. A high value resistor, R17, provides a DC path to ground for the capacitor. The resistor prevents the leakage currents through the diode clamp or rectifier diodes from charging C12, which would create a false logic 1 indication to U3 when operating from the battery. Schmitt trigger NAND gate U3A converts the analog 60-Hz signal from the low-pass filter into a digital signal that is used in the logic circuit. The other input to U3A selects the crystal oscillator 60-Hz signal for the clock time base by applying a logic 0.

Monostable U4 is the heart of the logic circuit since it makes the decision that controls the enabling of the crystal oscillator time base and the disabling of the display. The monostable is used as a missing pulse detector by making it retriggerable with a pulse width slightly longer than the period of the 60-Hz input. By making the monostable retriggerable, its $\overline{Q}$ output will stay a logic 0 as long as the line supply is providing 60-Hz signals, since the pulse width is longer than the input signal period. When line power is lost, the monostable times out and the $\overline{Q}$ output changes to a logic 1, which allows the crystal oscillator/divider output signal to toggle the output of U3B at a 60-Hz rate.

Since the accuracy of the monostable pulse width is important, a precision version of the regular CMOS dual monostable is used for U4. The device is specified to have a timing error of less than ±2 percent over its operating temperature range. To take advantage of the monostable accuracy the values of the two external timing components, C16 and R20, must also be accurate and stable. Metal or carbon film resistors, rather than composition resistors, should be used because they are available in 1 percent tolerances and are stable with respect to ambient temperature and moisture conditions. Polycarbonate or NPO ceramic capacitors are suitable for the timing capacitor application. Other types of capacitors using dielectric materials such as mylar, polyester, or Teflon® should be suitable in this application over the expected range of ambient room temperatures. General-purpose ceramics, as well as electrolytic and tantalum capacitors, should not be used for the timing capacitor because their values will change with temperature, time, and applied voltage.

Because the monostable pulse width is equal to the product of C16 and R20, a wide range of component values can be used. While the application literature from the different monostable producers regarding the range of component values is conflicting, the resistor

value can go from a minimum value of 5 kilohms to a maximum value of 10 megohms. The capacitor values are allowed to go from a minimum of 5000 pF to a maximum of 100 μF. For capacitor values above 0.5 μF, a diode from the timing resistor-capacitor junction to the supply voltage is recommended to protect the monostable. The minimum resistor value is determined by the current sink capabilities of the monostable, while the maximum value is limited by the leakage currents in the monostable and printed wiring board. The minimum capacitor value is limited by the parasitic capacitances associated with the monostable and the printed wiring board, while the maximum is limited by the current sink capabilities of the monostable.

Other components outside these two ranges will allow the monostable to operate, but the pulse width will differ from the formula value. The design pulse width for the monostable is 19.0 milliseconds, about 14 percent longer than the 16.7 millisecond period of a 60-Hz signal. The longer pulse width duration allows for a 3 percent error in monostable timing accuracy, a 2 percent error in timing resistor value and a 10 percent error in timing capacitor value. The component value ranges are wide enough to eliminate any parts procurement problems. If the exact values needed cannot be obtained, the monostable pulse width can be made longer than 19.0 milliseconds without having a significant effect on clock operation. The only drawback to increasing the monostable pulse width is that the dead time in the transition between line power time base and crystal oscillator time base will increase slightly. Designing the correct monostable pulse width was preferred instead of building the circuit then going through a time-consuming testing process to select the correct resistor value.

The remaining three Schmitt trigger NAND gates, U3B, U3C, and U3D, are for logic functions involving the operation of the time display and time-base selection. Ordinary NAND gates would be suitable in these three applications if the logic is changed for any reason. The design goal for the clock circuitry was to keep the number of parts to a minimum, so the logic was built to make do with the single quad NAND IC. Other logic modes can be included by the addition of different gates.

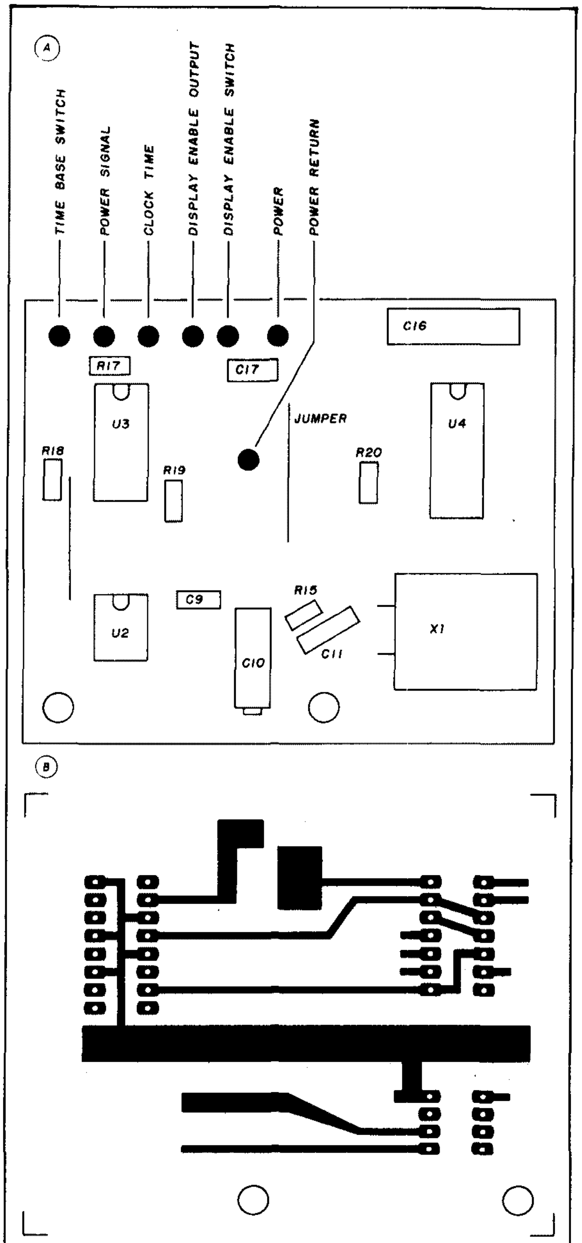

fig. 2. Digital clock PC logic board. (*A*) parts layout; (*B*) wiring pattern (actual size).

modifying clocks

With the MM5314 clock IC, modifying a clock or kit is very simple. Most of the work lies in mechanical areas. The locations of the new circuits, battery, and switches must be determined using the available clock volume and the existing installed components. The first step in the electrical modifications is to wire the battery with its blocking diode CR5 and changing resistor R1. The existing clock circuit board should have enough room to mount the diode and resistor, as well as numerous places to pick up a power return connection for the battery's negative lead. A push-pin drill and a No. 66 bit (0.031 inch/0.79 mm) can be used to make holes in the printed wiring board for new component leads. A small piece of copper foil with an adhesive backing can be used to make a junction point for the battery's positive lead, the resistor, and the diode.

table 2. Parts list.

component	description
B1	Nicad battery (GE No. GC-9B or Radio Shack No. 23-126)
C1	0.005 μF, 1 kV ceramic
C2	220 μF, 25 volt electrolytic
C3-C17	0.47 μF, 25 volt ceramic
C4-C8	0.01 μF, 25 volt ceramic
C9	0.1 μF, 25 volt ceramic
C10	4-26 pF variable
C11	33 pF, 25 volt mica
C12-C15	0.01 μF, 25 volt ceramic
C16	0.1 μF $\pm$10 percent polycarbonate (see text)
CR1-CR5	silicon rectifier diode (1N4002)
CR6-CR7	light emitting diode (MV50)
CR8	small signal silicon diode (1N914 or 1N4148)
F1	fuse 1/8 ampere
Q1-Q6	PNP switching transistor (2N3906 or 2N2907)
Q7-Q13	NPN switching transistor (2N3904 or 2N2222)
R1	7.5 kilohm (see text)
R2	470 kilohm
R3-R9	270 ohms
R10	4.7 kilohm
R11-R14	10 kilohm
R15	22 megohm
R16	100 kilohm
R17	1.0 megohm
R18-R19	10 kilohm
R20	191 kilohm $\pm$1 percent precision film resistor (see text)
S1-S4	single pole, single throw toggle switch
S2,S3,S6	single pole, single throw, normally open nonlatching pushbutton switch
S5	double pole, single throw toggle switch
T1	power transformer 12.6 volt/200 mA (Radio Shack No. 273-1385)
U1	MM5314 clock integrated circuit
U2	MM5369AA oscillator/divider
U3	CD4093B quad NAND Schmitt trigger
U4	MC14538B dual precision monostable
U5-U11	LED seven segment common cathode digit (FND No. 70, DL704, MAN 74 or Radio Shack No. 276-067)
V1	transient suppressor (GE No. V130LA10)
X1	color burst crystal, 3.58 MHz (Radio Shack No. 272-067)

After the battery has been charged, the battery backup mode can be tested by unplugging the line cord. If the clock shows the time it displayed when unplugged, the battery is doing its job. (The time display may dim, but this is expected because the battery voltage is lower than the line-supply voltage.) When plugged back in, all clock functions should return to normal. There should be no jumps in the time displayed by the clock as it's unplugged or plugged back in. The battery must be disconnected, and the clock unplugged, whenever additional modifications or changes are made to the clock.

Adding the oscillator/divider and logic involves three steps. The first is to connect the power and power return to the board, and connect the three-mode logic switches. The next step is to connect the display enable logic signal to pin 1 of the clock IC, U1. If this pin on the clock IC is floating, or connected to the supply with a pullup resistor, just connect the logic signal. When the pin is connected to the supply, the connection must be broken before connecting the logic signal. Operation of the logic circuit can be tested at this point before completing the modifications. Connecting the battery, and unplugging the clock, will power up the circuit. If the time display is on, changing the position of the time base switch, S7, should disable it. With the 60-Hz line power signal not yet connected, there is no input to trigger the monostable, so the logic should disable the time display. Once the display is off, pushing display enable switch, S6, should cause it to come on.

The last modification step is to disconnect the 60-Hz signal into pin 16 of the clock IC, U1. The clock should have a network similar to R16, C12, and CR8 to filter and clamp the input to the clock IC timing input. Break the connection of this network to pin 16, and connect the clock time signal from U3 pin 11 to U1 pin 16. Then connect the network junction to U3 pin 1 to complete the modification. The clock can now be set to the correct time, the crystal oscillator frequency set, and the remaining logic features tested.

construction details

A painted aluminum case houses the clock's electronics. A 3 × 6 × 4-inch (7.62 × 15.24 × 10.16 cm) deep case provides ample room for the three boards, battery, power transformer, and logic control switches. Smaller cases could be used, but mechanical layout and assembly would be more difficult. As shown in **fig. 1**, the three boards are divided by function. One board contains the clock integrated circuit, rectifier, and display driver circuits. The second board contains the LED digits and colon, with the last board containing the oscillator/divider and logic circuits. A complete parts list is provided in **table 2**. My clock was assembled from purchased clock and display circuit boards, so only the oscillator/divider, and logic components are shown, (**fig. 2**). The board is single-sided, copper-clad material. The copper layer is the top, or component side, and is used as the ground. Copper foil tape on the glass-epoxy side of the board is for wiring runs. The wiring layout was kept to one side of the board at the expense of using jumper wires for crossing runs. Mounting holes are at the bottom

of the board, with terminals at the top for wiring the board to the remainder of the clock circuits.

One unexpected problem was that the ambient room light made it difficult to read the time display at certain viewing angles and also made the display printed wiring board and the LED DIPs visible. The solution was to cut a piece of red acetate to fit behind the clear acrylic plastic windows in the case. The color of the acetate matches that of the LED display so that the digits can be read, but exterior light is filtered out. Commercially available display bezels use colored or smoked plastic covers to cut glare, but are more expensive than colored acetate.

operation

Once the clock has been wired, and the logic operation checked, there isn't much left to do. The oscillator frequency should be measured after a few months of operation and reset if aging has lowered the frequency. Only yearly frequency checks are necessary after that because the aging rate will reach its asymptote. The battery should be checked and inspected periodically for leakage or failure; the simplest way to check the battery is to unplug the clock and see if it works under battery power.

Since the clock is portable, it can be carried to a new location without resetting. But keep in mind that some of the components will not operate properly under extreme temperature conditions. Most commercial-grade ICs are specified for operation between 32 degrees F and 158 degrees F (0-70 degrees C) so if you take it out in the cold weather the clock should be insulated or heated. After setting the oscillator frequency of my own clock at work, where a very accurate calibrated frequency counter was available, I transported it home in 0 degree F (– 16 degrees C) weather. The clock IC held the time it displayed when I left the building because the cold stopped the crystal oscillator. Once back home, the oscillator warmed up and started, leaving me with a half-hour hole to account for. If outdoor or automotive operation of the clock is planned, the components should be either industrial (– 13 to + 185 F/ – 25 to + 85 C) or military (– 67 to + 257 F/ – 55 to + 125 C) temperature range rated to ensure proper operation.

conclusion

Adding a battery power backup and crystal oscillator time base to line-powered digital clocks is a practical way to overcome their operating problems when line power drops out. This clock has been running for a year and a half — including successful battery-powered operation through three long outages totaling 11 hours, and through plenty of shorter dropouts — without being more than 10 seconds off from WWV time.

ham radio

How does 0.1-905 MHz coverage sound to you?

add general coverage
to Yaesu's latest VHF-UHF receiver

The newly released Yaesu FRG-9600 receiver represents a significant advance in price/performance capability: it's the first communications grade receiver to offer all-mode capabilities, computer interface, frequency synthesis and broad coverage — 60-905 MHz. The receiver design is based on the use of already designed TV and FM modules and integrated circuits. These subassemblies have been combined in a very small package that gives good performance and makes for a highly flexible bench receiver.

There are few things, however, that one might wish had been done differently: the image rejection is modest and poses some problems in RF-dense urban areas. This problem is compounded by front-ends that have good sensitivity but poor large-signal handling capability. Band selection data is available at a connector on the back panel, but the VCO tuning voltage is not available. This makes the use of external tracking filters difficult to implement. Finally, incorporating one additional conversion to allow reception of the 0.1-60 MHz range would have resulted in a receiver with outstanding coverage and flexibility.

As one of the early users of this receiver, I decided to add an external HF converter that would cover the range 0.1-60 MHz, thereby *giving the receiver an overall range of 0.1 to 905 MHz.*

basic converter scheme

In the HF part of the spectrum, very large signals are available at the antenna terminals because of the size of the antennas used. It was determined, therefore, that a converter with good large-signal performance was essential. Image rejection problems can be easily handled by using an up-converter. (An up-converter simply means that the first IF is higher than the highest received frequency.)

In this case, the frequency range 0.1 to 60 MHz is converted to 100.1 to 160.0 MHz, permitting the display of the FRG-9600 to be read directly in MHz by simply ignoring the most significant digit (the 1 in 100 MHz). Unwanted mixing products are reduced about 20 dB by the use of a commercial double balanced mixer. Overall gain of the converter is about 0, which retains the original sensitivity characteristics of the receiver. Finally, all the antenna switching is automatic when the converter is activated. The basic converter block diagram is shown in **fig. 1**.

circuit description

The complete schematic diagram for the converter is shown in **fig. 2**. Power is provided from the auxiliary 8 volt RCA phono jack on the back panel of the FRG-9600. This jack can deliver 8 volts at 200 mA, more than enough for the converter. Power is applied through switch S1 and a protective diode. When the converter is off, relay K1 is not energized and the antenna connection from the VHF/UHF jack on the converter is straight through to the FRG-9600. When S1 is closed, the converter is activated and K1 is energized, switching to the HF antenna and the output of the up-converter.

Signals from the HF antenna jack are applied to a

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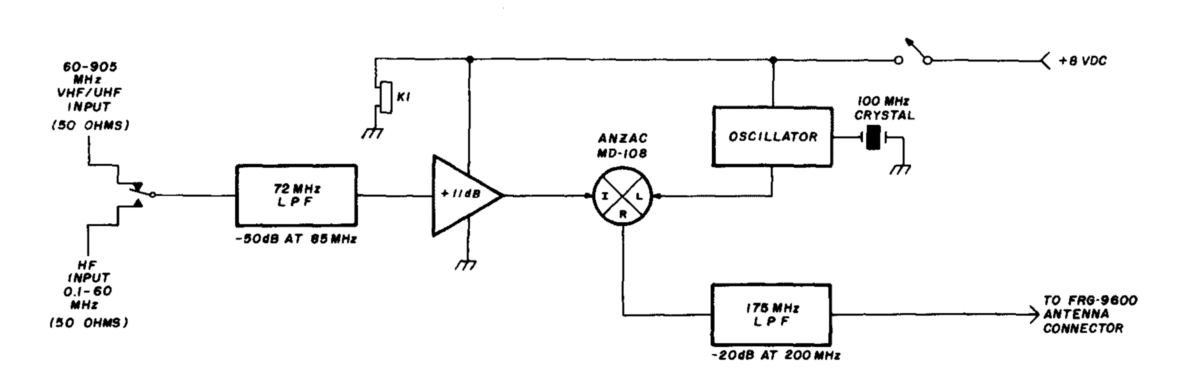

fig. 1. FRG-9600 HF converter block diagram.

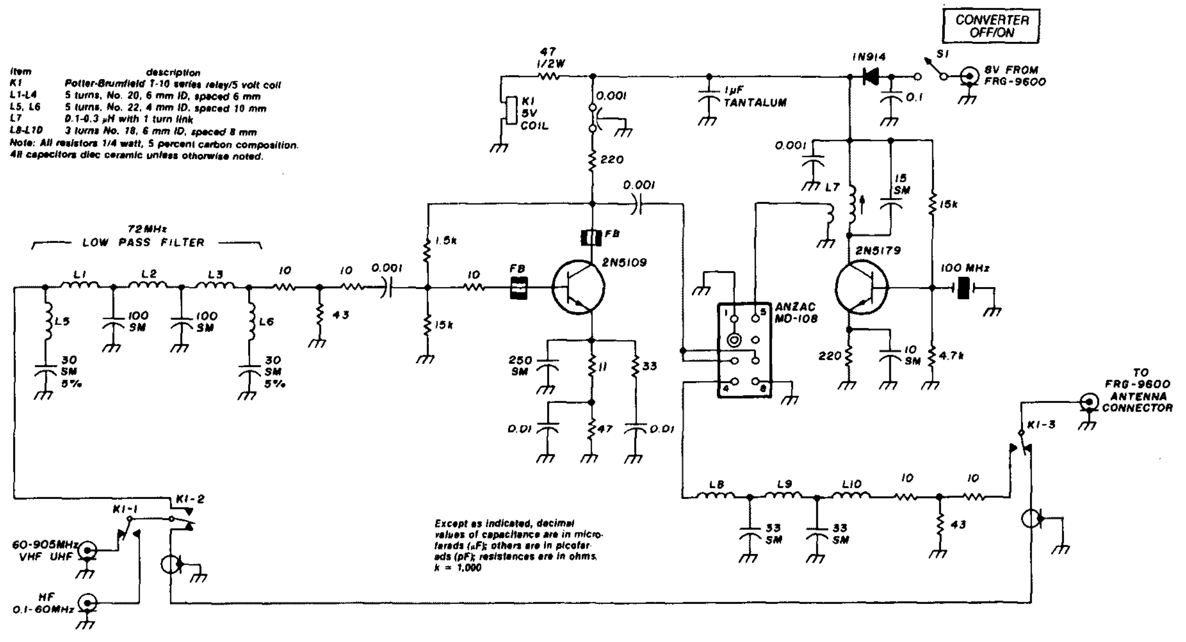

fig. 2. FRG-9600 HF converter 0.1-60 MHz.

low-pass filter and broadband large-signal amplifier. The amplifier stage uses a 2N5109, which is designed for low distortion, large-signal CATV applications. Note the rather complex network in the emitter of the 2N5109. This network shapes the overall response of the amplifier stage, making it quite flat from about 1-90 MHz. Gain of this stage is approximately 11 dB, and the noise figure is about 4 dB — more than adequate for general purpose HF work. The RF stage can handle input signals of approximately 0 dBm before any observable compression takes place.

The input filter is absolutely essential to the operation of the converter. Because the output of the converter is in the middle of the commercial FM band, it's important that FM signals through the RF amplifier be heavily suppressed. The input filter has a cutoff frequency of 72 MHz and is down 50 dB at 85 MHz. The filter should be built in close compliance to the design shown. The use of silver mica capacitors is a must, and they should have the shortest leads you can work with. Be sure that L2 is oriented 90 degrees to L1 and L3 so that no coupling takes place between any of the inductors. This same precaution applies to the inductors in the output filter. Note that these filters are

terminated by a simple resistive pad to assure that their characteristics are preserved with reactive terminations.

Because the 2N5109 is capable of considerable gain well into the VHF region, good VHF practice should be followed in layout and construction. All capacitor leads should be as short as possible. Ferrite beads on the base and collector leads suppress any tendency the stage might have to oscillate at VHF. The prototype showed no instability under any operating conditions. The mixer and local oscillator are straightforward. The 2N5179, used with a 100 MHz series resonant crystal will deliver about +8 dBm of LO power to the mixer — just right for the MD-108. This particular oscillator design is not very low-noise, but it's easy to use and adjust, and phase noise is not a major consideration in this application. Simply adjust the slug in L7 until the oscillator starts reliably each time. If you have an RF voltmeter, adjust L7 for maximum signal at pin 5 of the balanced mixer.

The Anzac MD-108* is a low cost general purpose mixer good to several hundred MHz. Other general purpose units should perform equally well. Since the mixer creates products and harmonics of the input signals, some precautions are in order. Because we're using the mixer to up-convert, the RF and IF ports are reversed from their normal order (see **fig. 1**). The output is taken at pin 4, and consists of some HF feedthrough, plus the mixing products of 100 MHz. The 175 MHz low-pass filter assures that we do not hear products that would otherwise be repeated, with diminished amplitude, every 100 MHz throughout the receiver tuning range.

construction and comments

Construction details are best left to each user. I used an LMB CR-800 enclosure that matches the appearance of the FRG-9600 nicely, but any *well-shielded* box will serve as an adequate enclosure. Construction on a piece of epoxy-glass circuit board material is simple and effective. Lay out the circuit more or less as the schematic is drawn, and you should have no problems. I strongly recommend using the exact component values shown for best results. Be sure that the cable from the converter to the FRG-9600 is high-quality coax to prevent signal leakage that would reduce the effectiveness of the low-pass filters.

The individual filters, active circuits, and the completed converter were characterized using a Tektronix 7L14/TR502 spectrum analyzer/tracking generator combination.

*Check pin designations carefully, several different configurations exist. RadioKit will provide a complete kit of parts, including PC board.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

first, the bad news

"How long will we have to wait until the sunspot cycle improves?" More and more Amateurs are migrating to the low frequency bands as the popular, higher-frequency DX bands are becoming more spotty. Ten meters is nearly deserted, except for an occasional north-south opening and some wandering ignition noise. And 15 meters isn't much better! Even 20 meters is a pale imitation of its former robust self.

Predicting when the sunspot minimum arrives and when the new sunspot cycle begins is a chancy business best left to the experts. A good guess indicates that sometime between winter, 1986, and spring, 1987, may be the turning point at which we move on to the next new cycle.

But several months of the new cycle must elapse before the high-frequency bands will come alive. Fall 1987 may be a good time to take the 10-meter beam out of mothballs and get it up in the air. That's two years away!

Meanwhile, there's a migration to the lower frequency bands and hams are turning to dipoles, inverted-Vs, delta loops and slopers. Certainly, some big DX "guns" have full-size 40 and 80 meter beams, but such monster antennas are out of the question for most operators.

now, the good news

Although we can't fool Mother Nature, there's still a lot of DX and good operating pleasure left on the "DC bands." As far as DX goes, many operators have made DXCC and won other juicy awards on both 40 and 80 meters. And I understand that Wal, W8LRL, has over 200 countries to his credit on 160 meters!

One of the better newsletters about 160-meter DX is published by Ivan Payne (VE3INQ). Send two IRCs and a business-sized envelope to Box 276, Station A, Weston, Ontario, Canada M9N 3M7 for this 22-page bulletin that will prove to you that DX is alive and well on 160 meters.

Along this line, Ivan's newsletter describes a simple 160-meter DX antenna, sketched in **fig. 1**, used at VS5RP by Bob Parkes (P29BR). Basically, it's a short, vertical antenna top-loaded by a single wire and inductively coupled by a toroid transformer to a coax line.

Bob recommends using from 25 to 40 radials. In his particular location, taking ground resistance into effect, he estimates the antenna's efficiency to be about 40 percent.

With regard to the radials few Amateurs can lay out 135-foot (40.7-meter) quarter-wavelength, 160-meter radials. The solution is to simply do the best you can. Several ground rods at the antenna feedpoint are useful, as well as a square of 1-inch (2.54 cm) mesh chicken wire laid on the ground. Dennis Peterson, N7CKD, uses a 30-foot square of chicken wire for a 160-meter ground screen plus other random ground connections to a metal fence.

The Canadian Top Band News also points out that long-path openings occur on the 160-meter band, citing the contact between AA1K (Delaware) and YB5AES (Indonesia) at 2205Z in October, 1984, as well as the contact between VE1ZZ (Nova Scotia) and 9M2AX (Malaysia) at 2323Z in January, 1985.

Finally, it should be pointed out that there's a 160-meter net active on Saturdays at 1600Z on 14.260 MHz and also on Tuesdays and Thursdays on 1840 kHz at 0400Z. DX and antennas are the main topics of conversation.

Speaking of antennas. . .

a very compact antenna for 160 meters

You can't get a full-size dipole up on 160 meters? You have a poor ground? You can't make a low resistance ground connection? Join the club! Most Amateurs have one or more of these problems. Unless you live in the middle of a large salt marsh, you're going to have to make compromises in your "top band" antenna system.

Some lucky Amateurs have enough space to squeeze in a large vertical antenna and lay out a number of radials. And others can erect loaded dipoles, or some form of Marconi antenna with a good ground system. But what about the rest of us?

A friend of mine wanted to get on 160 meters. He had about 55 × 25 feet (16.76 × 7.61 meters) in his backyard to work with, and his ground was terrible — rocky, sandy soil.

The only simple solution I saw was to erect a highly-loaded dipole antenna about 50 feet (15.2 meters) long. That would fit in the available space, and the dipole doesn't rely upon a ground connection to function properly. Such an antenna is shown in **fig. 2**.

The design is based upon a readily

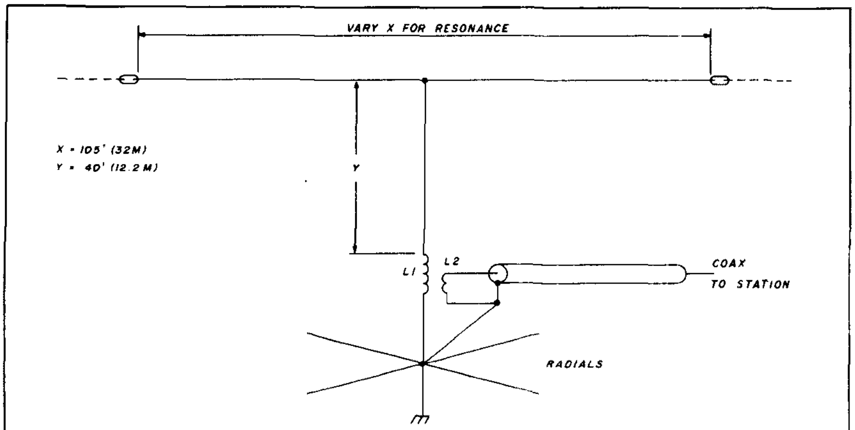

fig. 1. 160-meter antenna at VS5RP. L1, 28 turns No. 14 enamel; L2, 20 turns No. 14 enamel. Both L1 and L2 are wound on a 2-inch (5.08 cm) toroid, μ = 10 (Amidon T-200-2 or equivalent). Vary turns of L2 for best match.

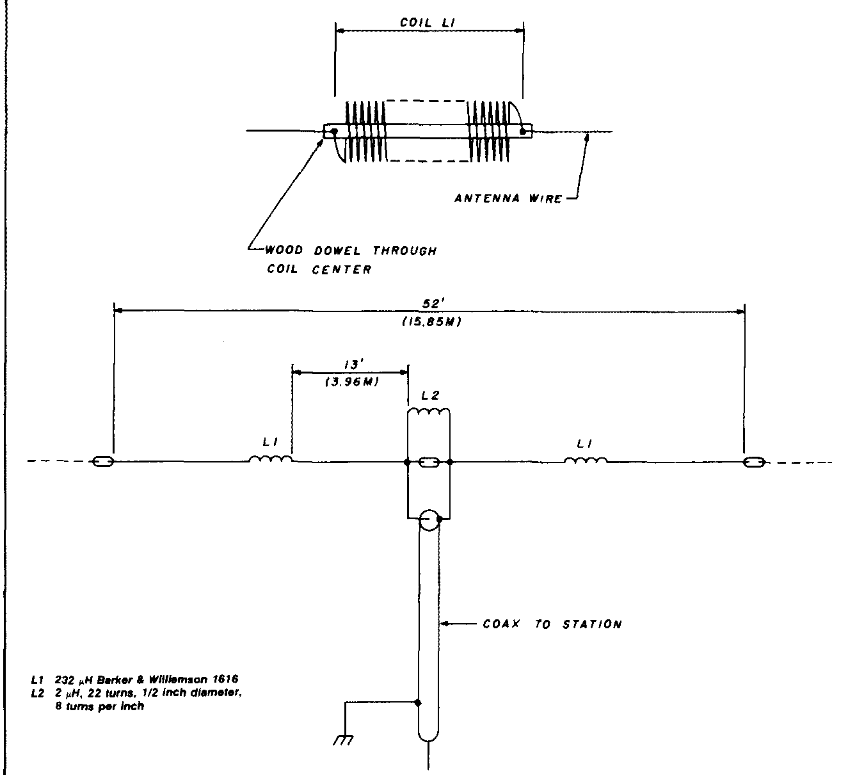

fig. 2. Assembly data for 160-meter "mini-dipole."

available, high efficiency loading coil: the Barker & Williamson* 1616 inductor. This coil is air-wound, 2 inches (5.08 cm) in a diameter and 10 inches (25.4 cm) long. It has 16 turns per inch of tinned copper wire. (It's also available with Formvar® coated wire, which should provide somewhat better efficiency than the tin plating when the coil is used in antenna service.)

Two of these ready-wound coils are used in this antenna, one in the middle of each leg. Since the coils are somewhat fragile, they're supported on an insulator made of a wood dowel rod cut to the same length as the coil. The ends of the antenna wires are passed through small holes drilled in the dowel, removing tension from the concentric coil.

The radiation resistance of the antenna is about 3 ohms, but the feedpoint resistance is close to 20 ohms, due to the loss of the coils. This results in an antenna efficiency of about 13 percent. This may make purists who have experienced little loss in their high-frequency antennas shudder, but the 160-meter band is a different matter and most of the small antennas used by Amateurs on this band exhibit a comparable degree of efficiency. The radiated signal, then, is about 8 dB down from that of a 100 percent efficient antenna (a dipole, for example).

A simple matching coil is placed at the center of the antenna to match it to a 50-ohm coax line. When properly adjusted, the antenna has a bandwidth of about 25 kHz between the 2:1 SWR points on the feedline.

antenna adjustment

The first step after building the antenna is to sling it up between two temporary points, allowing it to sag down until the center feedpoint can be safely reached from the top of a step ladder. The halves of the antenna are shunted with a two or three-turn link coupled to a dip oscillator. The resonant frequency of the antenna is carefully measured (with the aid of a calibrated receiver) and the antenna tip sections trimmed equally, a few inches at a time, until the antenna is resonant at your design frequency. (This one was cut for 1820 kHz.)

The pickup coil is removed and another coil is installed for matching to the coax feedline. The antenna is erected in its final operating position. The number of turns in the matching coil is then adjusted until unity SWR is obtained at some frequency near the design frequency. You'll find that the presence of the coil tends to detune the antenna a bit, and by the time

*Barker & Williamson, 10 Canal Street, Bristol, Pennsylvania 19007.

you've achieved a good match, the resonant frequency of the antenna will have moved.

The final step is to readjust the tip sections equally until the resonant frequency is back where you want it.

The whole process sounds tedious, but it's really not. The experimental antenna was built at an easy pace over one weekend and all adjustments were made during one morning of the following weekend.

And the antenna works fine! Granted, bandwidth of operation is restricted and antenna efficiency is low. However, running 150 watts input, contacts across the continent have been made on the band and, unless attention is drawn to the unusual antenna, most operators "on the other end" will assume you have a full-size dipole, judging from the reports my friend has received with his little antenna.

using an antenna tuner

Smart 160-meter operators know that a narrow-band antenna such as this compact dipole can be "pulled" in frequency by using an antenna tuner at the station end of the coax feedline. The very high off-resonance SWR exhibited by the antenna can be reduced to an acceptable value by the tuner. Experiments have shown that the antenna, with a simple tuner, permits operation over 100 kHz of the 160-meter band., And that's not bad for such a midget!

keep TVI to a minimum!

Two words of caution on this familiar topic: try *not* to run the antenna parallel to the house wiring system. It's easy to couple power from any 160-meter antenna into the house electrical wiring, but doing this can cause TVI, RFI, and other undesired reactions. In addition, since the coil loss of the antenna is high, don't try to run a lot of power into it. A good limiting figure for this antenna is 150 watts, so it will work OK with your exciter, but you'll burn up your antenna coils if you run your linear amplifier into it.

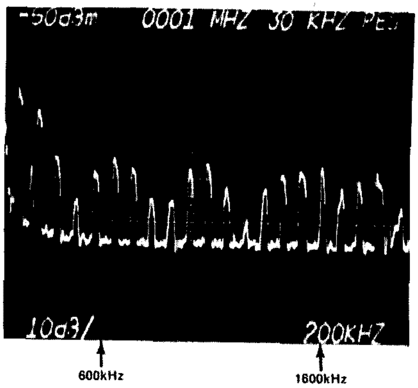

fig. 3. RF lighting device emissions (courtesy *Broadcast Engineering*).

I understand Barker & Williamson can supply coils with LEXAN® insulation instead of cellulose acetate or plexiglass. The extra cost of LEXAN is justified because it's impervious to the ultra-violet radiation from the sun that quickly destroys the plastic supports in the regular coils. A LEXAN-insulated coil wound with Formvar-coated wire sounds like the ideal inductor for any long-life loading coil exposed to the weather.

the RF light bulb

In my June, 1984, column I mentioned the possibility of RFI from the next generation of light bulbs. Although the subject lay dormant for months, the threat is real. In a recent issue of *Broadcast Engineering*, M.C. Rau, of the National Association of Broadcasters, wrote:

The pending introduction of RF lighting technology will significantly cut energy costs, by replacing the ubiquitous incandescent light bulb with RF devices. Unfortunately, many RF lighting devices emit energy at AM broadcast frequencies, both over the air and through the power line **[fig. 3]**. *A current FCC Notice of Inquiry is exploring the issues of lighting, the need for regulation of such equipment, and interference protections to be provided to the AM radio service.*

If RF lighting significantly increases interference over existing devices, NAB should act to ensure that the FCC adopts regulations carefully designed to protect the AM radio service.[1]

Well said! But a glance at the right-hand portion of the plot of **fig. 3** shows that RF emissions continue well above 1600 kHz, into the HF spectrum and probably the 160-meter and 80-meter Amateur bands.

It would be well for some enterprising Radio Amateurs who have appropriate facilities at hand to examine RF light bulbs, to see what problems they produce in the HF spectrum. NAB is doing a good job — as far as they go — but they have little interest above 1600 kHz. A word to the wise. . .

the 2-meter EME directory

I have additional copies of the 16-page 144 MHz EME Directory of active "moonbounce" participants compiled by Lance Collister, WA1JXN. You can obtain a copy by sending four first-class postage stamps (or four IRCs) to me at EIMAC, 301 Industrial Way, San Carlos, California 94070. (Don't send an envelope — we have oversize ones especially for this directory.)

reference

1. Michael A. Rau, "Charting a Course for A.M. Improvement," *Broadcast Engineering*, April, 1985.

ham radio

short circuit

tapered vertical

A misplaced parenthesis in "Calculating the Impedance of a Tapered Vertical" (K3OQF, August 1985, page 25) resulted in an incorrect calculation in **eq. 1**. The corrected equation should read as follows:

$$Z_0 = 60\,\ell n\left(\frac{2L}{b}\right) + 60\left(\frac{t}{b-t}\right)\cdot\ell n\left(\frac{t}{b}\right)$$

Upon substituting values on page 26, Step 1

$$Z_0 = 60\,\ell n\left(\frac{2\cdot 720}{1.5}\right) + 60\left(\frac{0.375}{1.5-0.375}\right)\cdot\ell n\left(\frac{0.375}{1.5}\right) = 384.3$$

All the other formulas and evaluations are correct. (TNX N6DH — Ed.)

VHF/UHF WORLD

Joe Reisert W1JR

transmission lines

When I first started this column almost two years ago, I made a list of the most important subjects that VHF/UHF/SHFers talk about. One was transmission lines, a subject I've only passively addressed in prior columns. Since I'm often asked to suggest a transmission line to someone, I thought this would be a good time to review the whole gamut of transmission line characteristics. Tradeoffs and data is presented so that you can select the optimum transmission lines for your applications.

Transmission line losses can be just as important as antenna gain, especially to the VHF/UHF/SHFer. If the same transmission line is used for both the receiver and transmitter, each dB of loss reduces the system capabilities by 2 dB (1 dB on receive sensitivity and 1 dB on transmitted power).

Often an antenna mounted preamplifier is used, especially on EME, to circumvent the received signal loss. This can be a costly and complex solution that helps only on the received signal.

transmission line types

There are many types of transmission lines. The most common are coaxial cable and balanced line on the lower VHF/UHF frequencies and waveguide on the SHF frequencies. Microstrip and stripline are types frequently used in low power and receiver type circuits. Lesser known or used types are the "G" line and Yagi types.

transmission line characteristics

Everyone knows that the purpose of a transmission line is to transfer power from one place to another. But most forget that a transmission line is nothing more than a low-pass filter. **Figure 1A** shows the equivalent circuit of a balanced line, while **fig. 1B** shows the same for an unbalanced line. If the values of L and C are known, the impedance can be determined as follows:

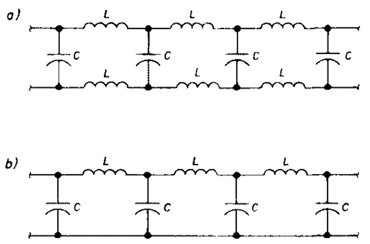

fig. 1. Representative circuits for a transmission line: (a) balanced line and (b) unbalanced, such as coaxial line.

$$Z = \sqrt{\frac{L}{C}} \qquad (1)$$

where Z is the characteristic impedance and L and C are the inductance in henries and capacitance in farads per unit length, respectively, of the transmission line. For example, if the L is 73 nanohenries and the capacitance is 29 pF for the same unit length (12 inches or 30.5 cm), the characteristic impedance is approximately 50 ohms. Terminating this line in 50 ohms will yield a 1:1 VSWR.

parallel balanced lines

This type of transmission line is often referred to as open wire or twinlead (**fig. 2A**). Well known and widely used in the past, it's seldom used nowadays except in stacking harnesses.[1] Open wire lines are usually inexpensive to make or purchase. Also, if properly constructed, the insertion loss is usually quite low.

The impedance of an open wire line is a function of the spacing of the wires, the dielectric between them, and the diameter of the conductors. The correct formula for determining the impedance of a symmetrical open wire line with air as a dielectric is:

$$Z_0 = 120 \cosh^{-1} \frac{b}{a} \qquad (2)$$

where a is the diameter of the conductors and b is the center-to-center spacing of the conductors in the same unit of measure as a (**fig. 2A**).

However, if the spacing is much greater than the conductor diameter (the usual case), a simpler formula can be used:

$$Z_0 = 276 \log_{10} \frac{2b}{a} \qquad (3)$$

From these formulas it is obvious that the most practical impedances are between 200-600 ohms. To prevent feedline radiation, the spacing should not exceed 0.05 wavelength at the frequency of operation. Despite rumors to the contrary, open wire lines do not radiate power even when the VSWR is high if the lines are properly constructed and kept well balanced.

On the negative side, balanced lines must be kept as straight as possible and away from nearby objects. Bends should be gradual, typically less than 45 degrees, and other adjacent lines or objects should be at least 2 to 3 times the width of the line away. This is a practical problem when multiple antennas or transmission lines are present and especially if a rotator is involved. Furthermore, open wire lines are often affected by moisture and insulator contamination, especially if the VSWR is high on the line. Therefore, the number of insulators should be kept at a minimum commensurate with maintaining proper spacing.

Twin lead is an acceptable type of balanced transmission line, but because it has a dielectric, it's typically lossier than open wire line. And unless you use the heavy duty type (such as

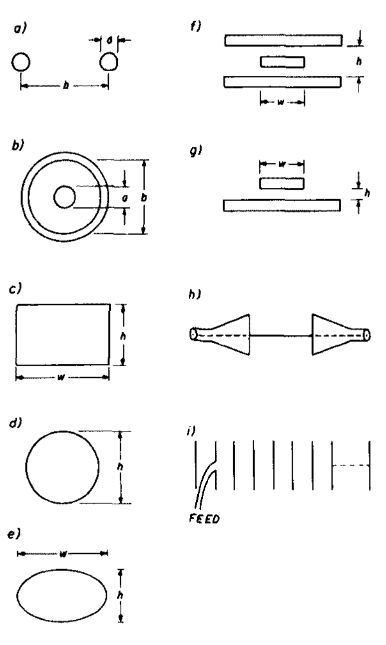

fig. 2. Typical transmission lines and their important dimensional parameters: (a) open wire line; (b) coaxial line; (c) rectangular waveguide; (d) circular waveguide; (e) elliptical waveguide; (f) stripline; (g) microstrip; (h) G line; (i) Yagi line.

K200), the power handling capability can be moderate to low. Furthermore, twin lead is often more affected by moisture than open wire line since it has a larger surface area.

coaxial transmission lines

Because of these problems with balanced lines, coaxial cable — in plentiful supply and in so many varieties — has become very popular. It can usually be bent and placed directly alongside other lines without the degradation associated with open wire or twin lead. The availability of accurate coaxial types of VSWR and power meters has further contributed to the use of this type of line.

Coaxial cables are often referred to as TEM (transverse electromagnetic) structures. They are an unbalanced type of transmission line. The impedance is a function of the dielectric as well as the ratio between the inside diameter of the outer conductor and the outside diameter of the inner conductor (fig. 2B). The impedance for an air dielectric coaxial transmission line can be readily calculated using the following equation:

$$Z_0 = 138 \log_{10} \frac{b}{a} \qquad (4)$$

where a is the outside diameter of the inner conductor and b is the inside diameter of the outer conductor in the same units of measurement (fig. 2B). For example, if the inner conductor diameter is 0.25 inch (6.35 mm) and the inside diameter of the outer conductor is 0.572 inch (14.6 mm), the characteristic impedance of the line will be approximately 50 ohms.

As with the spacing of the open wire line, the inside diameter of the outer conductor limits the upper frequency of operation. Roughly speaking, the upper frequency limit of a coaxial cable, the point at which other modes will propagate, is reached when the circumference of the inner diameter of the outer conductor is greater than approximately 1 wavelength at the frequency of interest.

waveguide

There was a time when waveguide was used as low as 200 MHz! It is by far the lowest-loss type of transmission line but is very costly and physically large below SHF. Waveguide can be either rectangular, circular, or elliptical (figs. 2C, D, and E).

Although the rectangular type is the most common, it's difficult to work with and often is referred to as "plumbing." The most common rectangular waveguide is used in the dominant or TE 1, 0 mode.[2] It's usually twice as wide as high and covers only about an octave in frequency. The width would be about 1/2 wavelength at the lowest usable frequency.

Elliptical semi-rigid waveguide is becoming more popular especially on microwave links since it can be moderately bent without distorting the characteristics. Circular waveguide is usually used in the below cutoff mode. It's very often found in precision attenuators and on the air outlets of high-power vacuum tube transmitters.[3]

Nowadays, coaxial cables are overtaking the use of waveguide as high as 26 GHz primarily because they're easier to work with. For those more interested in waveguides, reference 2 or any waveguide manual is recommended.

strip transmission lines

Microstrip and striplines are becoming very popular especially in receiver circuits. Often they are improperly identified by Amateurs. The true stripline is configured like a sandwich (fig. 2F). Note that it has a top and bottom, so essentially it is completely shielded.

Microstrip is by far the most common type of printed transmission line and is like an open-faced sandwich (fig. 2G). The field is not constricted to just the region between the strip and the substrate. Some field lines exist from the top of the strip to the substrate. Hence it is somewhat more difficult to design if there are tuned lines or adjacent circuitry. Other strip type of transmission line variations such as suspended substrate are also used, but they are beyond the scope of this month's column.

lesser known transmission lines

Up to this time I have been concentrating on the more popular types of transmission lines. The "G" line (fig. 2H) is a frequently overlooked transmission line that has some very interesting properties.[4,5,6] Originally called a surface wave transmission line, it was later named after its inventor, the late Dr. Georg Goubau, who designed the first such line in 1950. It resembles the "string telephones" that many of us made and used when we were children.

This type of transmission line operates in the TE mode. The launcher is like a large cone. The incoming line, usually a 50-ohm line, is impedance matched at each end to a single wire that travels the full length of the transmission line, primarily in the magnetic field.

If the wave is properly matched into the launcher at the input end, it will travel along the single wire to the com-

plimentary launcher at the opposite end. The attenuation will be extremely low provided that the launcher is properly fabricated, the impedance is matched and the correct wire is used.

Recently I had a chat with Warren Weldon W5DFU, who has done extensive work on a 23-cm tropo installation using this type of line.[7] He told me that the lowest loss line he constructed used No. 14 AWG (0.062 inch or 1.6 mm) Teflon™ covered 0.015 inch (0.38 mm) stranded wire that had only 1.2 dB per 100 feet (30.5 meters) of insertion loss at 23 cm! For those who can't afford expensive transmission lines on the UHF/SHF bands, this could be a real breakthrough.

At the ARRL National Convention in San Jose, California in 1965, Dr. Donald K. Reynolds, K7DBA, described a most interesting transmission line (**fig. 2**). It consisted of a Yagi-like structure analagous to the slow wave portion of a conventional long Yagi antenna. He indicated that if the proper spacing and element length were chosen (I believe he used 1/8 inch or 3.2 mm diameter wires approximately 0.4 wavelength long and spaced about 0.4 wavelength) that the line loss would be very low. He used a piece of flexible fiberglass to hold the rods. Although this type of structure could be frequency sensitive, it could have great potential, especially on monoband setups.

coaxial cable characteristics

So far in this column I've been talking in generalities, emphasizing that the coaxial cable type of transmission line is by far the most popular type at this time. Because this is true, I'll devote the rest of this column detailing other important things we should also know before we can select the optimum coaxial cable for our installations.

One of the first things that comes to mind is the dielectric. Typically speaking, air is the best dielectric because it has the lowest loss. However, this can be misleading. This is true only if the air is dry. Any moisture present will increase the loss dramatically.

One of the earliest coaxial cables was RG-8, which I believe was developed for radar installations during World War II. It turned out to have an impedance of 52 ohms. One of the advantages of a solid dielectric is that it is not likely to be affected by moisture.

The early dielectrics were principally made of polyethylene. Later the losses were decreased by using different types of foam. Some of the modern foams are almost as low-loss as air. But foam is not without its problems. Since it is usually softer than standard polyethylene, it can deform and even "cold flow." This is particularly true of Belden 8214 50-ohm line, so if you're using this type of line, don't bend it too sharply.

While on that subject, most coaxial cables have a minimum bend radius that should not be exceeded. It can be found on the manufacturers' data sheet, but as a rule of thumb, never bend a coaxial cable less than five and better yet ten times the cable diameter. For RG-8 type this would be approximately 2 to 14 inches (15 to 10 cm).

Some coaxial transmission lines are much better for bending. Rigid lines such as waveguides are not readily bendable. Don't be misled by the term "semirigid," because this type of cable is often very stiff and can usually be bent only once. Subsequent bends may break the outer shield. Generally the cable suitable for the jumper around a rotator are the ones that use a dielectric and braided shield such as RG-8/U and RG-213/U (more on this later).

attenuation and power

Probably the most important parameters when selecting a transmission line are insertion loss and power rating. Several factors affect these parameters, including impedance, size, dielectric, conductor material, and frequency of interest.

It is well known that the insertion loss of a transmission line is affected by the characteristic impedance. The lowest loss per unit length is between 180-220 ohms for open wire line, 70-75 ohms for microstrip and 75-80 ohms for coaxial cables. Coaxial transmission line loss versus impedance is shown in **fig. 3**. This is the principal reason the CATV industry uses 75-ohm impedances.

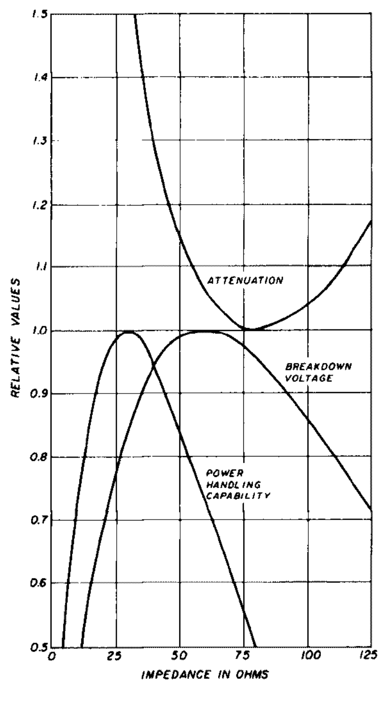

fig. 3. Some of the important parameters of a coaxial transmission line versus impedance such as attenuation, power handling, and voltage breakdown (adapted from reference 2).

Why isn't 75 ohms the Amateur standard, too? Well, there are other properties of coaxial transmission lines that also must be considered: power handling and voltage breakdown, for example. Note on **fig. 3** that the best power handling occurs in the vicinity of 30 ohms, while the best breakdown voltage is around 60 ohms. Hence the American standard of 50 ohms, a compromise between power and attenuation. (In some parts of Europe 60 ohms is the standard. Fifty ohms, however, is now becoming quite universally accepted, mainly because of all the test equipment and coaxial connectors available at that impedance.)

Material and size are also important: copper is the lowest-loss conductor;

aluminum has a higher loss but it is often used to minimize weight and cost. Frequently a copper-plated aluminum center conductor — such as found in the CATV industry — is used. The larger its physical size (as long as you don't exceed the cutoff frequency previously mentioned), the lower the loss and the higher the power handling capability. No wonder broadcast stations use such large transmission lines!

Coaxial transmission lines use many types of dielectrics. Using air as a dielectric can be an expensive proposition that often requires special connectors and a nitrogen pump to keep the air purged and hence non-contaminated.

Teflon™ dielectric cables such as the 0.141 numbered types of microwave semirigid coax are often used by Amateurs.[8] Recently RG-141/U and similar Telfon™ types of coaxial cable have become popular in situations where high power is required on a small diameter toroidal balun and on VHF/UHF antennas where a 1/2 wavelength balun is used.[9]

Modern foams are very low-loss dielectrics. This is particularly evident in the Belden 8214 and the newer 9914 RG-8 types of coax. The Andrew Corporation has introduced a dielectric called LDF (low dielectric foam) that rivals the loss of air lines. I'm sure that improvements will continue to be made as the requirements for smaller and lower loss coaxial cables increases.

Without a doubt, the most important question Amateurs are constantly asking is "Which coax is best?" This isn't an easy question to answer.

In order to help you decide, I prepared **table 1**. It lists most of the popular coaxial cables used by VHF/UHF Amateurs along with velocity factors. Insertion loss per 100 feet (30.5 meters) and maximum power ratings at 100 and 1000 MHz are also listed. These are manufacturers' typical ratings. A new coaxial cable in good condition should have an insertion loss that is equal to or less (but not much less!) than the figures shown.

You may ask what good the information is since it's specified only at 100 and 1000 MHz rather than on Amateur bands. Fortunately insertion loss increases at a somewhat logarithmic factor. To find the approximate insertion loss at a higher frequency for 100 feet (30.5 meters) of coaxial line, use the following straightforward equation:

$$I.L. = A\sqrt{\frac{F_H}{F_L}} \qquad (5)$$

where $I.L.$ is the actual insertion loss in dB, A is the attenuation in dB at the reference frequency (100 MHz in this case), F_H is the higher desired frequency and F_L is the lower or reference frequency (100 MHz). Total insertion loss per foot/meter is linear, so if you have half as much coaxial cable, the loss will be half that shown.

For example, if we want to determine the loss of 100 feet (30.5 meters) of RG-8/U coax at 432 MHz using **eq. 5**, it will calculate to be approximately 4.6 dB. (See **fig. 4** which relates insertion loss to frequency. A few representative coaxial cables are included.) If you know the loss of any coaxial cable at a specific frequency, all you have to do is place a dot on the graph where the frequency and loss are known. Then draw a line through the dot, parallel to the lines already shown. You now can determine the approximate loss at the frequency of your choice without any calculations at all.

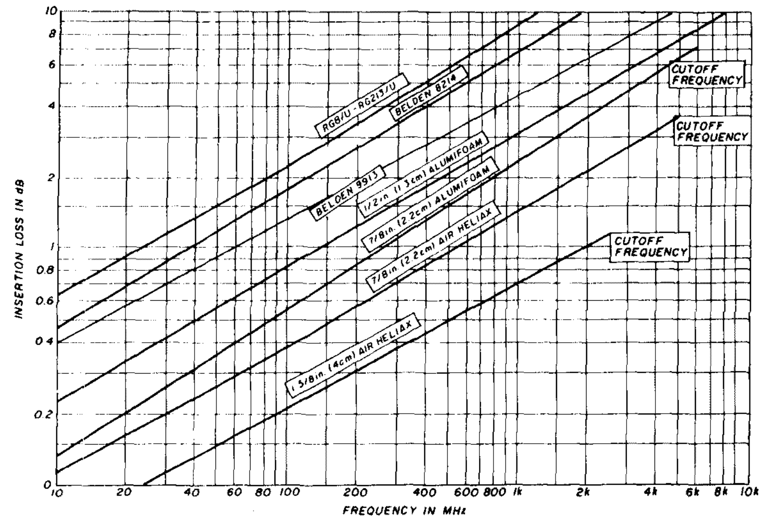

fig. 4. Attenuation versus frequency for 100 feet (30.5 meters) of some representative transmission lines. (See text for using this graph for other types of transmission line.)

Power handling is a more subjective rating, related to heating and breakdown. The rating can be approximated using the following equation:

$$P_X = P\sqrt{\frac{F_L}{F_H}} \qquad (6)$$

where P_X is the power rating at the desired frequency, P is the rating at a known low frequency, F_L and F_H is the desired higher frequency. For example, if we use the above example for RG-8/U, the power rating will be approximately 410 watts at 432 MHz. How many of you are exceeding this power level on 432 MHz?

Figure 5 shows power rating versus frequency for a few representative transmission lines. If you have a known power rating for another transmission line at a particular frequency, mark it on the graph, draw a line through the mark and parallel to those already shown, then read off the ratings at the desired frequency.

trade names

Some commercial transmission line

trade names are quite popular and often misused. The most common are Alumifoam™ and Heliax™. Alumifoam (often referred to as hardline) refers to a coaxial cable with a seamless aluminum outer shield and a low-loss foamed polyethylene dielectric. This type of feedline is quite common, especially in CATV systems.

Heliax is the trademark of the Andrew Corporation. In its early days, it resembled a helical or spiral corrugated outer shield, usually made of copper. The inner dielectric was either air or foam polyethylene and the center conductor was usually a copper wire or tubing.

In recent years the Andrew Corporation redesigned the foam type of Heliax and now uses an "annular" or concentric ring type of corrugation. This construction technique is supposedly less prone to moisture damage, since water seeping into the feedline would have more difficulty traveling down its walls than it would in the helical form of construction.

More recently the Andrew Corporation introduced an aluminum outer shield coax similar to the annular constructed line, but lower in cost and aimed mainly at the TVRO market. However, it has higher insertion loss than the standard copper shielded LDF types.

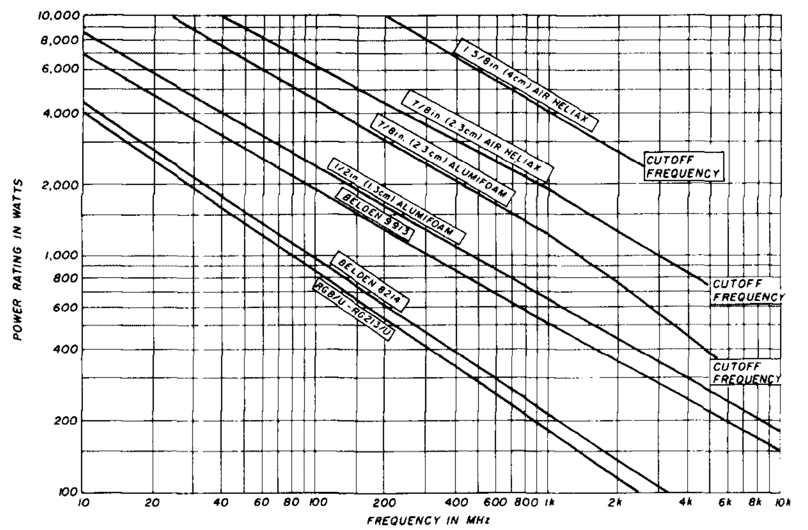

fig. 5. Power handling versus frequency for some representative transmission lines. (See text for using this graph for other types of transmission line.)

other mechanical considerations

The most common coaxial transmission lines are the braided and the solid shield. As suppliers have tried to lower costs, however, the number of strands in the braid has decreased remarkedly; as this decreases, so does the strength of the connection at the connector, resulting in increased insertion losses and decreasing shield effectiveness. At the same time, some of the better shielded types such as RG-214 have been priced right out of the Amateur market. In an attempt to offset these problems, some suppliers have introduced foil shielding, usually backed up by a few small strands of wire; although these changes are quite worthwhile, the shields are often hard to secure properly.

As mentioned before, flexible types of lines are recommended wherever bending or movement, such as around a rotator, are expected. Although these types usually have higher insertion loss, the Andrew Corporation has developed a special type of line called Superflexible Heliax,™ which has half the insertion loss of RG-8 or RG-213. In addition to the corrugated outer conductor, it has a stranded center conductor.

Most transmission lines can be buried without fear of water entry if they're free from nicks or holes in the outer sheath. The best types to bury are those that have a non-metallic outer protective jacket.

There's always the problem of getting the transmission lines into the shack. I use an ordinary clothes dryer vent and pass the feedlines right on through, stuffing an old rag into the rear of the vent to keep out rodents and lessen air flow.

Finally, there's the age-old question of contaminating versus non-contaminating jackets used on the typical polyethylene coaxial cables. Most RG-8 coax uses a contaminating jacket while RG-213/U doesn't. *Avoid the contaminating types of jackets at all cost. They're often slightly cheaper cables, but are not MIL SPEC. They start to deteriorate immediately, with sharp increases in insertion loss that can become disastrous in a few years. If you're not sure, don't buy it. If it's cheap, it's probably the contaminating type.*

connectors

Unfortunately, time and space will permit only a short discussion on this subject. I prefer type "N" connectors because they exhibit a good VSWR and reasonable power handling capability up to 500 MHz. *UHF connectors should be avoided at all cost, especially above 150 MHz. Not only are they poor on VSWR, but they often let in moisture.*

Hardline and Heliax types of coaxial cable require special connectors that are usually rather expensive but are necessary to preserve both the VSWR and integrity of the connection. This is particularly true of the air dielectric types. Failure to use the proper connector could allow moisture to enter and literally destroy the transmission line. Anyhow, why fight it? If

table 1. Typical characteristics of the most commonly used transmission lines by Amateur VHF/UHF/SHF'ers. Unless otherwise shown, all are 50 ohm types.

cable type	insertion loss in dB per 100 feet (30.5 meters) (see note 1) 100 MHz	1000 MHz	power handling in watts 100 MHz	1000 MHz	velocity of propagation
RG-58C/U	4.90	20.0	170	44	0.659
0.141 semirigid	3.60	11.6	2200	600	0.750
RG-8/U (note 2)	2.20	9.0	850	190	0.659
RG-213/U	2.20	9.0	850	190	0.659
Belden 8214	1.80	7.0	950	215	0.780
Belden 9914	1.60	6.0	1000*	250*	0.780
Belden 9913	1.40	4.5	1900	520	0.840
1/2-inch (1.3-cm) Heliax RG-268, RG-366/U	0.85	2.9	2200	570	0.790
1/2-inch (1.3-cm) Alumifoam RG-231/U, RG-331/U	0.82	3.1	2300	650	0.800
RG-17/U	0.80	3.8	3200	560	0.659
1/2-inch (1.3-cm) Air Heliax	0.80	2.7	2200	620	0.914
1/2-inch (1.3-cm) LDF Heliax	0.72	2.4	1900	530	0.880
7/8-inch (2.2 cm) Alumifoam RG-332/U, RG-333/U	0.55	2.3	4500	1250	0.800
3/4-inch (1.9-cm) ohm CATV	0.50	1.7	3300	950	0.800
7/8-inch (2.2-cm) Heliax RG-323/U, RG-324	0.50	2.1	4700	1200	0.790
1-inch (2.5-cm) 75-ohm CATV	0.40	1.4	4600	1200	0.800
7/8-inch (2.2-cm) LDF Heliax	0.39	1.4	5100	1400	0.890
7/8-inch (2.2-cm) Air Heliax	0.38	1.4	6100	1900	0.916
1-5/8-inch (4-cm) Heliax	0.30	1.4	9300	2000	0.790
1-5/8-inch (4-cm) LDF Heliax	0.23	0.9	14000	3500	0.880
1-5/8-inch (4-cm) Air Heliax	0.21	0.7	15000	4000	0.921

*Estimate.

Note 1. These are approximate maximum loss numbers for good quality new coax. In the case of air dielectric, these figures only apply if the cable is moisture free and is pressurized with dry air or nitrogen.

Note 2. The RG-8/U produced in recent years may have higher loss than noted.

you're using expensive line, use the proper connections and you won't lose what you've just gained!

Belden 9913 coax is becoming quite popular. However, the connectors are just becoming available and are quite expensive. In the meantime, UG-21B or Kings 59-207 connectors are recommended because they have an extra large and wide clamp that provides the necessary holding for the special foil and braid. In either case, the center pin of the connector has to be slightly enlarged with a drill, or the center wire of the coax must be filed down slightly, to gain access. The latter is *recommended.*

measurements

Always test each transmission line properly before installation. It's much easier to do this on the ground than after installation on a tower! Testing should be done with a VSWR/power meter such as the Bird Model 43 or equivalent. If you don't own one, plan to buy one and try to borrow one for the tests.

First connect the line under test to a good dummy load. Then place the VSWR/power meter between the output of a suitable transmitter and the transmission line to be tested. Next measure the VSWR looking into the line. It should be very low (typically less than 1.2:1 if the dummy load and the line are good).

Now measure the power going into the line under test. Then bypass the meter and place it at the load and again read the power. The difference in indicated power represents the insertion loss of the transmission line. If it's greater than the manufacturer's specifications, don't install it until you find the problem. After all, you don't want a dummy load between the shack and the antenna system!

Insertion loss always makes the VSWR of a load look better than it really is. In fact, a transmission line with a 10-dB insertion loss will indicate a 1.2:1 or better VSWR even if the line isn't terminated.[10]

If your feedline is lossy, how can you properly evaluate the VSWR of the antenna at the other end? The answer is that you can't unless you know the exact line loss and can calculate backward to determine the true load VSWR as described in reference 10.

Finally, if the VSWR on a transmission line is high, there's an additional "mismatch loss" over and above the feedline insertion loss.[11] General-

ly, if the VSWR is less than 3:1 at the load and the transmission line insertion loss is 3 dB maximum, this additional loss will be less than 1 dB. But who want to throw away any more hard to obtain dBs?

recommendations

We're now ready to make the final selection for our transmission line. First review the comments made earlier in this article and then study table **1** and **figs. 4** and **5**. If possible, try to obtain manufacturer's specifications. Most transmission line manufacturers have extensive catalogs with all kinds of information about their products.

If a long transmission line is needed, a combination of types is permissible. For instance, hardline or Heliax can be used from the shack to the top of the tower and from the antenna down the mast joined around the rotator with a flexible type of line. Other combinations are also acceptable providing that the overall insertion loss of the sytem is within reason.

Power lost in transmission lines is gone forever. Although some installations use antenna-mounted preamplifers to lower the received signal loss, this does not make up for the transmitted signal loss. After all, who needs to heat up Mother Nature?

This brings up an important but subtle issue. A 3 dB insertion loss transmission line would require 3 dB more antenna gain to offset the loss. This would mean at least doubling the antenna size while halving beamwidth and probably more than doubling the wind load. This type of problem can usually be partially solved by using a larger, albeit more expensive, transmission line.

For example, at 432 MHz 100 feet (30.5 meters) of RG-8 or RG-213/U coax would have a loss of about 4.6 dB, as mentioned earlier. Replacing such a line with the same length of 7/8-inch (22-mm) hardline would lower the loss to 1.15 dB, a drop of almost 3.5 dB! The total cost of this change would probably be only an additional $50.00, much less than the cost of increasing the antenna gain by 3 dB! *Don't be penny wise and pound foolish. Larger low-loss feedlines will pay handsome dividends and more than offset their initial cost by the performance gain.*

A further cost savings idea is suggested. Many of us install one expensive low-loss transmission line for UHF/SHF and place a remotely activated coaxial relay at the top of the tower. The number of antennas that can be accessed is only limited by the number of poles on the relay.

Suitable relays are often found at flea markets at attractive prices ($25-50), especially if you count in the number of poles. In this configuration the actual cost of the transmission line and relay is divided by the number of antennas that are accessible. For example, even using only an ordinary two-position coaxial relay effectively almost halves the cost of the transmission line per band without sacrificing performance!

Talking to old timers can clue you into some of the problems inherent in the use of certain coaxial cables. RG-17 is such an example. On a long run (perhaps 100 feet or 30.5 meters), the coefficient of expansion of the inner conductor and the outer conductor can be quite different. If the temperature rises or falls considerably, the center conductor may expand or contract more than the outer jacket. The net result can be a broken center pin at the interface when heated or a retracted center pin that breaks contact during cold weather. Use an "LC" type connector since the standard N connector, a UG-167, tends to increase the severity of this problem.

If you opt to use CATV transmission line, proceed with caution; some 75-ohm connectors are available, but the coaxial types are usually scarce. If you use this type of line, make an impedance transformer at least at the shack end to get back to 50 ohms to match your VSWR meter, etc.[12] And beware — CATV transmission line, especially the type that Amateurs usually obtain, is typically specified up to 350 MHz. Often the VSWR will go out of specification just above this frequency although it may improve at some higher frequency. *Caveat emptor.* Test it at the frequency of interest before you install it!

I'll let you in on a secret. I often find and purchase 50-ohm hardline and Heliax at Amateur flea markets. Sure, the price may be higher (50 cents to a dollar per foot) than conventional transmission line and connectors more difficult to find (although sometimes they're included on the line), but look at the reduction in price from the manufacturer and the difference in performance! One line like this will give you at least 10 to 20 years of uninterrupted superior performance. Can the lower cost cable do this? I'll bet you'll spend more in the same period of time by replacing the cheaper brand every few years — without the performance advantage.

summary

I've covered plenty of miscellaneous material in this month's column. But a thorough understanding of the information presented in the test, graphs, and table will give you most of the tools necessary to select the optimum cable for your requirements.

So there you have it. Just remember that the primary guideline when selecting a suitable transmission line is *don't be penny wise and pound foolish.*

acknowlegements

Many thanks to Warren Weldon, W5DFU, for bringing the information on the G-line to my attention.

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important VHF/UHF events:

October 5-6:	*Mid-Atlantic States VHF Conference, Warminster, Pennsylvania (contact WA2OMY)*
October 5-6:	*International Region 1 UHF/SHF Contest*
October 9:	*Peak of Draconids Meteor Shower predicted at 0300 UTC*
October 15:	*EME perigee*
October 20:	*Peak of Orionids Meteor Shower predicted at 1100 UTC*
November 2:	*Peak of Taurids Meteor Shower predicted at 0930 UTC*
November 2-3:	*ARRL International EME Contest*
November 3:	*Peak of Casseopids Meteor Shower predicted at 0930 UTC*
November 12:	*EME perigee*
November 17:	*Peak of Leonids Meteor Shower predicted at 0300 UTC*
November 23-24:	*ARRL International EME Contest*

ham radio

repairing flood damage

After four days of constant, heavy rains *the river crested 11 feet above flood stage and overwhelmed the best efforts of hundreds of bone-tired volunteers. Despite backbreaking, heroic efforts, the sandbag wall at the edge of town gave way under the relentless pressure of the angry river.*

Over the next 24 hours the water at your home QTH rose, completely flooded the basement, and gushed into the first floor to a height of 6 feet. Time was too precious to save anything but the family — all else was left behind. As the waters receded, the governor called out the National Guard to prevent looting, and you returned to recover what you could. You found your rig all but ruined, still damp and covered with mud.

Although most flood damage scenarios are not as dramatic as this one, we nonetheless hear of radio equipment that has, for one reason or another, taken a bath. Boating accidents, plumbing failures (Gee! Was that plastic pipe *really* running just above my radio set?) and a variety of other problems splash our rigs out of service. Fortunately, if the insurance company pays off well enough, you can go out and buy a new rig. But if the insurance company refuses to pay — "Sorry... wind-driven water damage excluded ..." — or if you don't have insurance, then you might want to take restorative action yourself. Even if the insurance company does pay, you can usually buy the rig back from them for salvage value. One guy I know received $325 for a two-year old transceiver and bought it back from the insurance company for $20. The company sent him a check for $305, and he kept the carcass.

Some of the steps I'll recommend may sound a little bizarre to you in terms of safety and comfort, but make more sense when you're faced with the possible loss of an expensive piece of equipment. Some of the steps — especially those involving baking the moisture out or using chemicals to clean the rig — might actually cause a little damage that will also have to be repaired. If *that* makes you nervous, it may help to remember that you cannot harm the radio any more: *it's already a total loss*! Any restoration is, therefore, pure gravy.

don't touch that dial!

The first thing to do is *refrain from turning the rig on*, even for a brief test to see whether it will or won't work. Satisfy yourself right now that even a short dunk causes fatal damage! Still, the all-too-natural urge is to see whether your rig survived the flood...*if it was immersed, then it didn't survive*!

So if your rig has been under water, remove the covers and give it a bath.

A bath?

I once lived in a seaport town where saltwater damage to electronic equipment was common. The shop where I worked part-time (while attending college) took in an $1800 UHF-FM taxicab radiotelephone set that had been immersed the night before during a storm; it seems that the saltwater river tributary had overflowed its banks enough to cover the radio, mounted in the trunk well. The first thing the shop owner did was take the transceiver out to the parking lot and give it a ten-minute shower with a garden hose. He'd lived in that town all his life, and had much experience with water-damaged radio gear.

(If the damage to your unit is due to saltwater, then do the cleaning job *immediately*. Don't delay; the longer salt residue remains in the equipment, the greater the corrosion damage will be.)

In some cases, it will be necessary to follow the shower with actual immersion. A friend of mine uses a 25-gallon tub, the kind you might use to bathe a large dog. In the tub, he mixes two to four quarts of a product like Lestoil, a small bottle (2 to 4 fluid ounces) of either fingernail polish remover or acetone (same stuff), and enough tap water to fill the tub all the way to the rim. He leaves the set in the bath for an hour, then pours out the solution and rinses the tub out thoroughly, refilling it with plain tap water. (Some people prefer to use distilled water, which is available in bottles in some areas). This second bath removes any residue left by the chemicals in the first bath.

Note: this Lestoil/acetone bath may damage some plastics. If this worries you, then use plain soapy water. It isn't quite as effective as solvent, but it works somewhat. Keep in mind that the damage will usually not prevent the rig from operating, and most plastic pieces can be replaced anyway. The rig is already a total loss, so don't worry about trivial secondary damage!

The next step is drying the unit out *thoroughly*. If you live in Arizona (yes, they have floods in the desert!), then simply leave the rig out in the sun for about a week. Everyone else will have to use some other method. The kitchen oven is a good bet, provided that it can be regulated to maintain a temperature of 125 to 130 degrees Fahrenheit (52-54 degrees Celsius). That range is low for a kitchen oven, and some ovens might not be able to remain that cool. Higher temperatures will dry the rig out faster, but will also melt some of the plastics used in the radio, so beware. The drying process takes several days — perhaps as long as a week.

Another way is to build a cardboard (or other material) box and use several hundred watts of incandescent lamps to provide heat. Use a thermometer in-

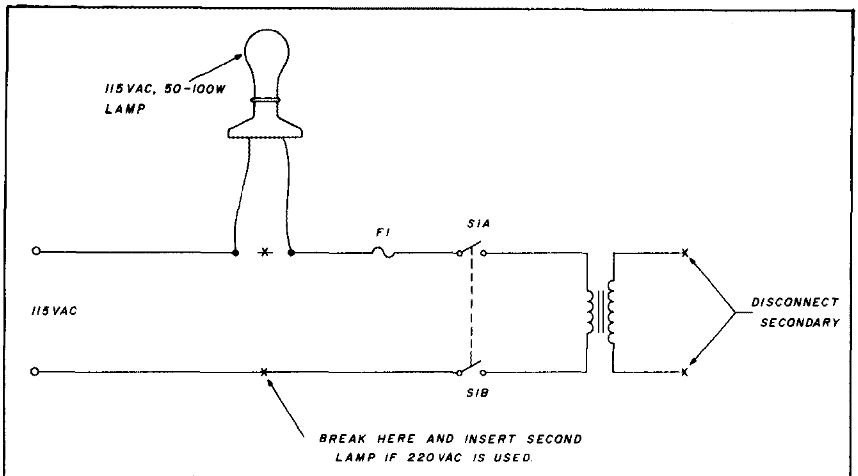


fig. 1. Method for driving moisture out of a high-voltage transformer.

side the enclosure to ensure that the 130 degree "melt limit" is not exceeded, and that the box doesn't catch fire. Again, as much as a week may be needed, although I have dried out a car radio that was dropped into fresh water (for a few minutes) in only one day.

preventing secondary damage

Now comes the big test! In some cases, the only way to test the equipment is to turn it on and look for smoke. I prefer a more conservative approach that sneaks up on it one step at a time. First I disconnect the internal DC power supply; this can be absolutely essential to the survival of the equipment being repaired, especially those with high voltage power supplies, such as certain transceivers and most linear power amplifiers.

Without connecting the rig to AC power, connect a bench power supply to the circuitry that was previously connected to the rig's internal power supply. It's essential to use a DC power supply that will provide the same voltage(s) as the original internal supply, and additionally (*this is important*) has a current limiter control. The output voltage is set to the DC voltage normally supplied by the rig power supply, and the current limiter control is set for a short-circuit current only a little above the normal operating current of the circuit under test.

Why go to such trouble? Because you want to prevent secondary damage. There's almost inevitably a short circuit or other condition that draws loads of current. If such a condition exists in the equipment, the internal power supply normally used will probably produce enough current to burn up components, printed wiring board tracks and other components. After the circuit is checked out, then we can check out the power supply and, if it's working, reconnect it.

The low-voltage DC power supply should be checked out separately, especially if it uses a series-pass regulator — almost all do these days. If the regulator circuit is not working, then one possible fault allows the rectifier output to be connected to the regulator output; this occurs when the series-pass transistor is either shorted or hard biased to full turn on. Since the rectifier voltage is always higher than the regulator output voltage, it can damage the circuits that were just pronounced healthy.

High voltage power supplies present special problems. Small amounts of moisture that are no problem in low voltage supplies will zap a high voltage supply into Never-Never Land. The special problem is the high-voltage transformer; extra drying may help, but if moisture has entered, then it may have to be replaced. **Figure 1** shows a method of drying a power transformer. A 115 volt AC lamp is placed in series with the primary of the high-voltage transformer. The current flow is sufficient to cause internal heat buildup, but not enough to zap the transformer if it is shorted. If the high-voltage power supply uses a 220 VAC primary circuit, one lamp should be placed in series with each AC hot line (see **fig. 1**).

Some remaining areas of concern, and probable damage, are those components where moisture can enter and remain hidden. Candidates include trimmer capacitors, air variable capacitors, IF and RF transformers, switches and potentiometers, paper capacitors, and electrolytic capacitors.

On trimmer capacitors, we can open the capacitor up to the minimum capacitance position (with the screw all the way out) and apply heat from a hair dryer or incandescent lamp for 10 or 15 minutes. Whether or not this step is necessary can be determined after the initial power-on test reveals a specific problem. Otherwise, you'll mess up the alignment of the rig for nothing. This step should not, therefore, be used merely as a matter of course, but only in response to a specific symptom.

Similarly, air variable capacitors may have corroded contact wipers between the rotor and stator, and this will be apparent when the rig is turned on.

Paper and electrolytic capacitors can absorb water, especially if they have a fiber or cardboard end cap. If the capacitor shows signs of being soggy, then replace it; capacitors are, after all, relatively inexpensive.

If a lot of scum remains on the printed circuit board, then spray clean it with Freon TF or a similar product. Use a small paint brush or a piece of cheesecloth to help loosen the scum.

Flood-damaged radios are often salvageable. However strange these methods may sound, they've been used successfully by professional service technicians for many years.

If you have a question, let me know. While I can't guarantee a personal answer, I will attempt to answer as many of your questions as possible in this column.

ham radio

PTFE* VHF antenna insulators

Turn your own for best performance and appearance

Several years ago I undertook a project that involved the construction of a number of identical antennas for 144 MHz and 432 MHz. Through-the-boom insulated elements seemed to be the best method of construction, but I was not satisfied with the insulators available. Because I'm retired and have a fairly well equipped home machine shop, I decided to make my own.

After experimenting with a variety of materials, and after many weeks of work, I finally discovered that Teflon™ had all the properties I was seeking but one — low price. Teflon possesses a happy combination of ductility and elasticity, which, when combined with its superior insulation property, makes it ideal for my purposes. When installed in the boom, it will lock itself to the boom with a friction fit on the element holding it firmly in place.

The machine-made insulator is much faster and easier to make and install than the hand-made version. Its dimensions are shown in **fig.** 1; the body of the insulator, measuring 0.312 inch (0.79 cm) diameter at the shoulder, tapers about 0.005 inch (0.013 cm) toward the other end. They're made to fit 5/16 inch (0.794 cm) holes (through the boom) and 3/16 inch (0.476 cm) diameter elements. The hole through the center is 1/8 inch (0.318 cm). When placed in the 5/16 inch (0.794 cm) hole in the boom and expanded, the insulator forms an internal shoulder that locks it to the boom.

lathe-turned insulators

My first insulators were made from 3/8 inch (0.953 cm) diameter Teflon rod. With the Teflon held in the headstock chuck, a hole was drilled through the center with a 5/32 inch (0.397 cm) inch drill. A series of cuts was made using a 3/16 inch (0.476 cm) wide chisel-

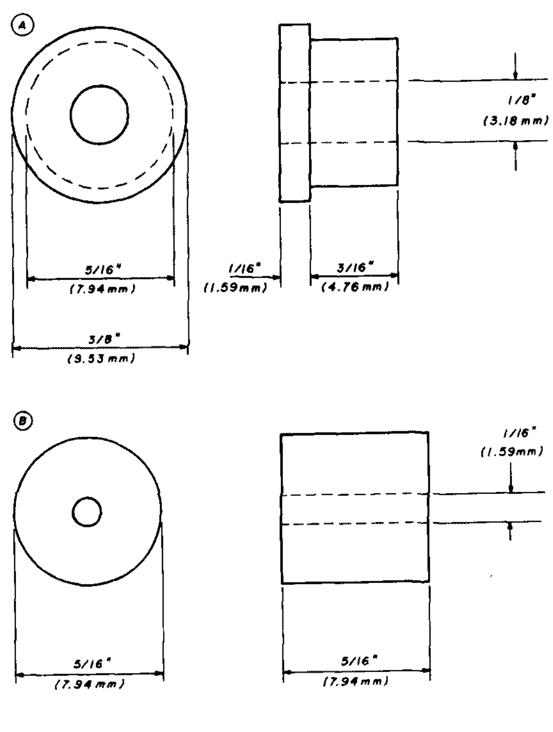


fig. 1. Teflon insulators: (*A*) machine-made, (*B*) hand-made.

type cutting tool, leaving a 5/16 inch (0.794 cm) inch diameter body and 1/16 × 3/8 inch (0.159 × 0.95 cm) shoulder. The machining was done with the drill bit remaining in the Teflon to support it and keep it rigid. The individual insulators were then cut apart with an Xacto™ knife held against the shoulder with the stock rotating in the lathe. In this manner I could make about eight insulators at one time.

There was one problem, however. Because Teflon of this diameter is quite flexible, when the drill extends into the Teflon that lies beyond the support of the chuck jaws, it tends to "wander" and become eccen-

By George Chaney, W5JTL, 218 Katherine Drive, Vicksburg, Mississippi 39180

*Polytetrafluorethylene

tric. At the time I was tapering the ends of the elements and simply driving them through the insulators, letting them expand in the boom. This approach worked quite well, and I still have some antennas in use that were assembled in this manner several years ago. I did not realize that Teflon would tolerate, without fracture, the degree of expansion I later discovered. Further research in my plastics supply catalog revealed that heavy-wall Teflon tubing is available (at almost double the price of the rod) in 3/8-inch OD and 1/8-inch ID. I bought some, tried it, and have used it as my basic material since then. With no holes to drill, my production rate skyrocketed. What had been a chore now became a pleasure.

The lathe configuration is illustrated in fig. 2. The heavy wall tubing is inserted through the chuck into the lathe spindle, extending approximately 2-1/2 inches (5.127 cm) out of the chuck. To support the Teflon during machining, use 1/8 inch (0.3175 cm) diameter drill rod, held in the tailstock chuck. Insert it in the center hole of the teflon tubing all the way to the headstock chuck. The drill rod acts as a mandrel to support the Teflon during turning and cutting processes, so there's no problem making ten insulators at one time. When finished they're simply removed from the mandrel, and the process is repeated.

It's not necessary to measure the shoulders for thickness. An "eye-ball" 1/16 inch (0.16 cm) is satisfactory. But it *is* necessary to measure the body and produce the insulators uniformly. I use a dial indicator caliper for this purpose. After turning a few, I take note of the cross-feed index. If I find that I'm getting uniform production, I put on a cross-feed stop, adjust it so that the cutting tool feeds into the work to the proper depth and can go no further. Thereafter, the cross-feed is fed in until it stops; it is then withdrawn, and the lathe carriage is moved forward 1/4 inch (0.635 cm). The process is repeated until the mandrel is full of insulators. (A skilled operator could easily turn out 200, and perhaps 300, insulators per hour.)

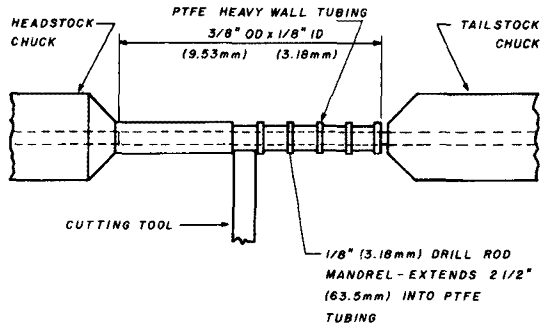


fig. 2. Metal lathe set-up for production of insulators. The cutting tool (chisel-type, 3/16-inch, 0.4763 cm) wide is fed straight into the PTFE to reduce diameter to 0.312 inch (0.792 cm). Shoulders are 1/16-inch (0.1588-cm) thick. (The drawing is expanded.) The clearance between the end of the Teflon and tail-stock chuck need not be more than 1/8-inch (0.3175 cm). After ten cuts have been made, insulators are separated by holding a sharp knife edge against the insulator "body" at the right-hand edge of each shoulder, with the work rotating in the lathe.

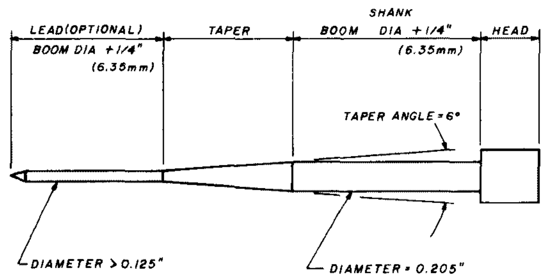


fig. 3. The expanding tool may be made of brass, bronze, aluminum, or steel. W5JTL uses a 1/4-inch (0.6350-cm) brazing rod.

installation

Insulators with a 1/8 inch (0.318 cm) center hole will require more than a slight taper of the element if the element is to be used for expansion. After much experimentation, I've concluded that a 6-degree included angle (3 degrees each side of center) taper is about optimum. This would result in reduction of the element diameter of nearly 1 inch (2.54 cm) at each end, and could adversely affect the design resonance, particularly at 432 MHz and above. My first expansion tools were 0.188 inch (0.476 cm) in diameter and less tapered — perhaps only 8 or 10 degrees; these occasionally produced sheared insulators. At the 1984 Central States VHF (CCSVHF) Conference, Jan King, W3GEY, expressed an interest in obtaining some of my insulators. I took him to my hotel room to give a demonstration of how to use them and — you guessed it — promptly sheared off a couple of them. Nevertheless, he left with a few hundred insulators and I came home and went back to the drawing board. I now make the expanding tools 0.205 inch (0.521 cm) in diameter, with a long taper. (These are illustrated in the drawings of fig. 3.) When withdrawn from the insulator, the hole immediately shrinks to 0.175 to 0.180 inch diameter and provides ample friction to hold the element.

Two opposite side insulators may be installed in the boom at one time if the shank of the installing tool is long enough to go all the way through and the bottom side insulator is supported until it is expanded.

A "lead" section on the expander helps in alignment, but is not necessary. A length of wood measuring approximately 2 × 4 × 12 inches (5.08 × 10.16 × 30.48 cm) with a hole large enough to clear the expander, drilled about 3 inches (7.62 cm) deep [centered on the 2 inch (5.08 cm) side] is a valuable aid in insulator and element installation. It provides support for the bottom insulator during this process. The boom is placed on the wood with the two opposite side insulators in place. The expander is driven through both of them and extracted. The element — with the sharp corners at the end is rounded off 0.005 inch (0.013 cm) with a file or sandpaper — is then driven through until it protrudes an inch or two on the opposite side. Inspect it from the bottom side to make sure that it's centered in the bottom insulator before driving it all the way through. Centering the elements in the boom is done after all have been installed.

hand-made insulators

If you're making a single antenna and want only a few dozen insulators, they can be made without a lathe. Unfortunately, the heavy wall tubing (and it's not available as a stock item with a center hole smaller than 1/8 inch, or 0.318 cm) will not expand sufficiently to form its own shoulders and "stay put" when driving the elements through. Teflon rod of 5/16 inch (0.794 cm) diameter is readily available and is much less expensive than either 3/8 inch (0.953 cm) rod or tubing. All you have to do is drill a hole through the center of it and slice it off into individual insulators. This is easily done in a drill press. Place a short piece of straight metal rod 5/16 inch (0.794 cm) diameter in the drill press chuck. Put a drill press vise on the rod and tighten it with the rod in the vertical "vee" of the vise. You can now drill short pieces, up to about 1-1/2 inches (3.81 cm) long, through the center with sufficient accuracy for our purposes. Use a drill no larger than 3/32 inch (0.238 cm) and preferably 1/16 inch (0.156 cm). Cut the individual insulators about 5/16 inch (0.794 cm) long.

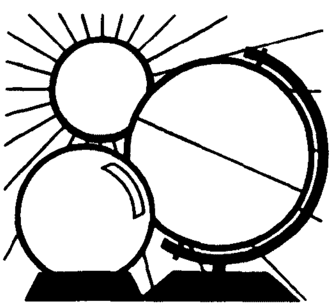
Installation is somewhat similar to the lathe turned insulators, except that they must be put in "bottom side" first. The expanding tool must have a point small enough to enter the smaller hole. Since greater forces are required in this installation, better support of boom and insulator is necessary. The insulator should be "half in and half out" of the hole in the boom during expansion. To maintain things in this position, a relief hole for the insulator is provided by fixing a piece of flat thin gauge metal (1/16 inch, or 0.159 cm, aluminum is OK) to the wooden block before drilling the hole for the expander. Then put a piece of 1/8 inch (0.318 cm) thick aluminum, with a 3/8 inch (0.953 cm) diameter hole through it, over the other metal piece, with the 3/8 inch (0.953 cm) relief hole centered over the expander hole. It can be held in place with glue. Place the boom with insulator in place, on the wooden block and centered over the 3/8 inch (0.925 cm) relief hole. Insert the expander through the vacant top hole in the boom and drive it through the insulator, expanding it. Withdraw the expander. Turn the boom over 180 degrees with insulator in place in what is now the bottom side, insert the expander through the previously installed insulator, and drive it through. It's now ready for element installation, in the same manner as the lathe turned insulators.

conclusion

I've made and disposed of several thousand of these insulators. Every user I've heard from has expressed complete satisfaction. The material cost for the hand-made insulators should be no more than 4 cents each, if quantity price of Teflon is obtained. I've made and will continue to make the machine-made variety available to VHFers at that price.

Perhaps some one else can produce them more economically. I claim no proprietary rights and invite anyone so inclined to produce them; I'll be glad to furnish more detailed information to anyone wishing to produce them commercially.

ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

sunspot cycle update

It's time for a six-month update on the progress of the 11-year solar cycle, and time for a look at what conditions might lie ahead. Could we be approaching the minimum?

First, consider the sunspot number (SSN) itself. Since our last update,* a slight increase of about 10 SSNs appeared in May 1985 and continued into June and July before decreasing to the present level.

One clue that indicates a SSN minimum is close occurs when sunspots of the new cycle appear on the sun's disk. For about two years the new spots are seen simultaneously with the old cycle's spots. This sighting of the new spots at high solar latitude, 30 to 35 degrees, and of opposite polarity to those in that hemisphere, occurs about a year before the minimum. This had not yet happened as this issue went to press.

Trends in the SSN 11-year cycle duration from minimum to minimum and the values of SSN at the minima are interesting to note. The cycles with higher values at their peak tend to be of shorter duration and have higher valued SSN minima following them (the peak). The present cycle 21 had a maximum monthly SSN of 165 (53 percent above the average) and consequently should end up being a short cycle (9.0 to 13.6 years, average = 11.1 years). The SSN minimum value (0 to 11.2, average = 5.1) should be higher than the average of 5.1. As of this month, the cycle is 9.3 years long, which is short compared to recent cycles, which varied from 10.17 to 11.4 years duration from minimum to minimum. Therefore, it's probable that only another eight to ten months will pass before we reach the SSN minimum; this puts the date of the SSN minimum somewhere between April and August, 1986. So the monthly average values are expected to decline slowly from the current 12 to about 6 or 7 by late summer of 1986.

Now let's look at the 10 cm solar flux as a predictor. Solar flux monthly averages decreased to 73.5 in October, 1984, and stayed within ±2 of 75 until May, 1985. Since flux is a direct energy measurement it is closely correlated with ionospheric effects. Also, daily values of flux can be used directly without smoothing; thus they are made available to us easily and quickly. In monthly averages of daily values the solar variation throughout the month is mostly eliminated because of the similarity of the lengths of the solar rotation (27 days) and a month (28-31 days). One value per month makes seasonal and annual solar effects easy to study. May, June, and July averages were up about five units, marking the return of greater 27-day cycle activity. The flux average was raised more by the activity than it could decrease since the decrease was limited, being so close to the lowest flux ever recorded (63) for a day. The lowest value of flux so far in cycle 21 was 68 on May 31, June 1, and June 26, 1985. The solar flux monthly average is expected to slowly rise or at least remain constant through the winter before decreasing those few remaining units after spring into summer. Solar flux minima tend to occur during summer when the sun is furthest away; the 27-day solar variation is often less in summer months. It is interesting to note that if the 27-day variation is absent the daily values come close to the monthly average near SSN minimum.

*DX Forecaster, *ham radio*, April, 1985, page 84.

last-minute forecast

The higher HF bands are expected to be very good after the 12th and during the third week of October. It's probable that the solar flux 27-day maximum (as small as it is these days) will occur about that time. That, added to the beginning of the rise in solar flux to winter levels as the sun-earth distance shortens, may bring a good maximum. The higher band openings should be the result of long-skip transequatorial propagation, particularly if moderate geomagnetic disturbances appear at that time. Geomagnetic disturbances are most likely about October 21 to 27. The lower frequency bands 30 to 160 meters, should greatly improve as a result of decreased thunderstorm noise and lower attenuation. Both should provide increased DX range in the evenings. These lower bands should be best the third and fourth weeks of the month.

The Orionids meteor shower will be visible from the 15th to 24th of October, with a maximum rate of between 10 to 20 per hour on the 20th to 21st of the month. The moon is full on the 28th and perigee occurs on the 15th. A total eclipse of the moon on the 28th begins at 1515 UT in the Western Pacific along the countries of New Zealand, Eastern Asia and part of the Arctic. It will travel across the Indian Ocean, Africa, and Europe, ending in Iceland and Eastern Greenland at 1929 UT.

band-by-band summary

Ten, twelve, fifteen, and twenty meters will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of these bands will be shorter and will occur closer to local noon. Transequatorial propagation on these bands will more likely occur toward evening during conditions of

OCTOBER

WESTERN USA

GMT	PDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	5:00	20	40	20	10	12	12	10	20
0100	6:00	20	40	20	10	12	12*	10	20
0200	7:00	20	40	20	12	12	12*	10	20
0300	8:00	30	40	20	15	12	12*	15	20
0400	9:00	30	*40*	20	15	15	12	15	30
0500	10:00	30	*40*	30	20	15	12	15	30
0600	11:00	40	*40*	30	20	20	15	15	40
0700	12:00	40	*40*	*30*	20	20	15	20	40
0800	1:00	*40*	*40*	*30*	*20*	*20*	20	*20*	40
0900	2:00	*40*	*40*	*30*	*20*	*20*	20	*20*	40
1000	3:00	*40*	*40*	*30*	*20*	*30*	*20*	*20*	*40*
1100	4:00	*40*	30	*30*	*20*	*30*	*20*	*20*	*40*
1200	5:00	*40*	30	15	*30*	*30*	*20*	*20*	*40*
1300	6:00	*40*	20	12	20	*30*	*30*	*20*	*40*
1400	7:00	30	20	10	12	*20*	*30*	*30*	*40*
1500	8:00	40	20	10	10	20	*30*	*30*	*40*
1600	9:00	40	20	10	10	15	*20*	*30*	*40*
1700	10:00	40	20	10	10	15	20	20	*40*
1800	11:00	*40*	20	10	10	12	20	20	40
1900	12:00	*40*	20	10	10	12	15	30*	20
2000	1:00	*40*	30	12	10	12	15	20	20
2100	2:00	40	30	15	10	12	15	20	20
2200	3:00	40	40	15	10	12	12	15	20
2300	4:00	30	40	20	10	12	12	12	15

MID USA

GMT	MDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CDT
0000	6:00	20	40	20	10	12	12	10	20	7:00
0100	7:00	30*	40	20	12	12	12	12	20	8:00
0200	8:00	30	40	20	12	12	12	15	30	9:00
0300	9:00	30	40	20	15	15	12	15	30	10:00
0400	10:00	30	*40*	20	15	15	15	20	40	11:00
0500	11:00	40	*40*	30	20	20	15	20	40	12:00
0600	12:00	40	*40*	30	20	20	20	20	40	1:00
0700	1:00	*40*	*40*	*30*	*20*	20	*30*	20	40	2:00
0800	2:00	*40*	*40*	*30*	*20*	*30*	*30*	*20*	40	3:00
0900	3:00	*40*	*40*	*30*	*20*	*30*	*30*	*30*	40	4:00
1000	4:00	*40*	30	*30*	*20*	*30*	*30*	*30*	*40*	5:00
1100	5:00	*40*	20	20	*20*	*30*	*30*	*30*	*40*	6:00
1200	6:00	30	20	15	15	*30*	*30*	*20*	*40*	7:00
1300	7:00	30	20	12	15	*20*	*20*	*20*	*40*	8:00
1400	8:00	30	20	10	12	20	*30*	*20*	*40*	9:00
1500	9:00	30	20	10	10	15	*30*	*20*	*40*	10:00
1600	10:00	40	20	10	10	15	*20*	*20*	*40*	11:00
1700	11:00	40	20	10	10	15	20	20	*40*	12:00
1800	12:00	*40*	20	10	10	12	20	20	30	1:00
1900	1:00	*40*	30	10	10	12	15	20	20	2:00
2000	2:00	*40*	30	12	10	12	15	15	20	3:00
2100	3:00	*40*	30	15	10	12	12	12	20	4:00
2200	4:00	*40*	40	15	10	12	12	12*	20	5:00
2300	5:00	40	40	20	10	12	12	12	20	6:00

EASTERN USA

GMT	EDT	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	8:00	30	40	20	12	12	12	12	20
0100	9:00	30	40	20	12	15	12	15	30
0200	10:00	40	40	20	15	15	12	20	30
0300	11:00	40	40	20	20	15	15	20	40*
0400	12:00	40	*40*	20	20	20	15	20	40
0500	1:00	*40*	*40*	*30*	20	20	15	20	40
0600	2:00	*40*	*40*	*30*	*20*	20	20	20	40
0700	3:00	*40*	*40*	*30*	*20*	*30*	20	*20*	40
0800	4:00	*40*	*40*	*30*	*20*	*30*	20	*20*	*40*
0900	5:00	*40*	*30*	20	*20*	*30*	20	*30*	*40*
1000	6:00	20	20	12	*20*	*30*	*20*	*30*	*40*
1100	7:00	20	20	10	15	*20*	*30*	*30*	40
1200	8:00	20	20	10	15	*20*	*30*	*20*	40
1300	9:00	20	20	10	12	20	*30*	*20*	*40*
1400	10:00	30	20	10	12*	15	*30*	*20*	*40*
1500	11:00	30	20	10	10	15	*30*	*20*	*40*
1600	12:00	40	20	10	10	15	*20*	*20*	*40*
1700	1:00	40	20	10	10	15*	20	*20*	*40*
1800	2:00	40	20	10	10	12	20	20	*40*
1900	3:00	*40*	30	10	10	12	15	20	30
2000	4:00	*40*	30	12	10	12	15	15	20
2100	5:00	*40*	30	15	10	12	12	12	20
2200	6:00	*40*	40	15	10	12	12	12*	20
2300	7:00	*40*	40	20	10	12	12	10	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

highest solar flux and a disturbed geomagnetic field.

Thirty and forty meters will be useful almost 24 hours a day. Daytime conditions will resemble those on 20 meters. Skip distances and signal strength may decrease during midday on those days that coincide with the higher solar flux values. Nighttime DX will be good except after days of high MUF conditions and geomagnetic disturbances. Look for DX from unusual places on east, north, and west paths during this time. The usable distance is expected to be somewhat less than 20 in daytime and greater than 80 at night.

Eighty and one-sixty meters will exhibit short-skip propagation during daylight hours and lengthen for DX at dusk. These bands follow the darkness, opening to the east just before your sunset, swinging more to the south near midnight, and ending up in the Pacific areas during the hour or so before dawn on the path of your interest. The 160-meter band opens later and ends earlier than 80.

ham radio

a portable 2-meter beam

Briefcase antenna packs small, travels light

One recent Saturday morning, the topic at our ham club breakfast turned to the difficulty of keeping antennas up through Oklahoma's typical springtime season of frequent thunderstorms and occasional tornadoes. What we all wanted and needed, we decided, was a compact, truly portable 2-meter antenna that could provide reliable performance in virtually any setting.

In response to this challenge I developed an antenna that weighs less than a pound and stores in a space measuring less than 12 by 1-1/2 inches (30.5 × 3.81 cm). I chose a four-element Yagi beam because it is directional, can have vertical or horizontal polarization, and deliver respectable gain. Assemby and disassembly can be very quick. Once assembled, it can be hung from a tree in a campsite, suspended by fishing line and hooks in drapery and ceiling tiles, or hung in the living room, even during severe thunderstorms.

construction

The boom is constructed of a 45-inch (114 cm) piece of 3/8-inch (0.95-cm) OD thin wall aluminum tubing cut into four equal sections of 11-1/4 inch (28.6 cm) each. To join the boom sections together for assembly, make three pieces of solid aluminum 2 inches (5.1 cm) long and slide each inside the sections of boom tubing (see **fig. 1**) to restore the boom to its full 45 inch (114 cm) length. Where the boom sections join, measure 1/2 inch away from each side of the cut and drill a clearance hole vertically through the tubing and solid pieces to accommodate a 4-40 screw. A 4-40 × 3/4-inch (1.9 cm) screw is used to tie each section together. A standard 4-40 hex nut and washer are used on part of the assembly and washer and wing nuts to allow for quick assembly and disassembly on the other part (see **fig. 2**). When taken apart, the boom stores the reflector and director; the 4-40 screws and wing nuts hold the elements inside when stored.

Next turn the boom 90 degrees from the top/bottom plane to make the holes for the elements. Measure 3/4 inch (1.9 cm) from one end; this will be the location of the first element. Measure from this point 14-1/2 inches (36.83 cm) to the next point. Continue till all four element locations are marked. Then drill clearance holes for 4-40 screws perpendicular to the vertical holes for bolting the boom together. Next cut the heads off four 4-40 × 1-1/4 inch (3.18 cm) long screws. Insert them into the holes and secure with hex nuts and washers. The elements will be attached with these mounting screws.

The gamma match is made from a 6 inch (15.24 cm) piece of 1/4-inch (0.635 cm) tubing, a 6 inch (15.24 cm) piece of No. 14 insulated wire, a BNC connector and several brackets. First cut a piece of aluminum stock 1/2 × 1-1/2 inches (1.27 × 3.81 cm). Drill a clearance hole for the BNC connector. (Use a UG-1094/U connector.) Bend the stock into an L shape 1/2 × 1 inch (1.27 × 2.54 cm). Mount the connector in the bracket as shown in **fig. 1**. The shorting strap for holding the 6 inch (15.24 cm) piece of tubing to the driven element and tubing were made from aluminum stock. (On the first model, copper proved very satisfactory.) Solder one end of the No. 14 wire to the BNC connector and slide the tubing for the gamma match over it. Measure the stock for the shorting bar and bend and mount it to the driven element and 6 inch (15.24 cm) piece of tubing covering the gamma match wire; because the tubing will need to be moved during tuning, leave it slightly loose.

By John Eighmy, KB5QJ, 511 East 14th Street, Bartlesville, Oklahoma 74003

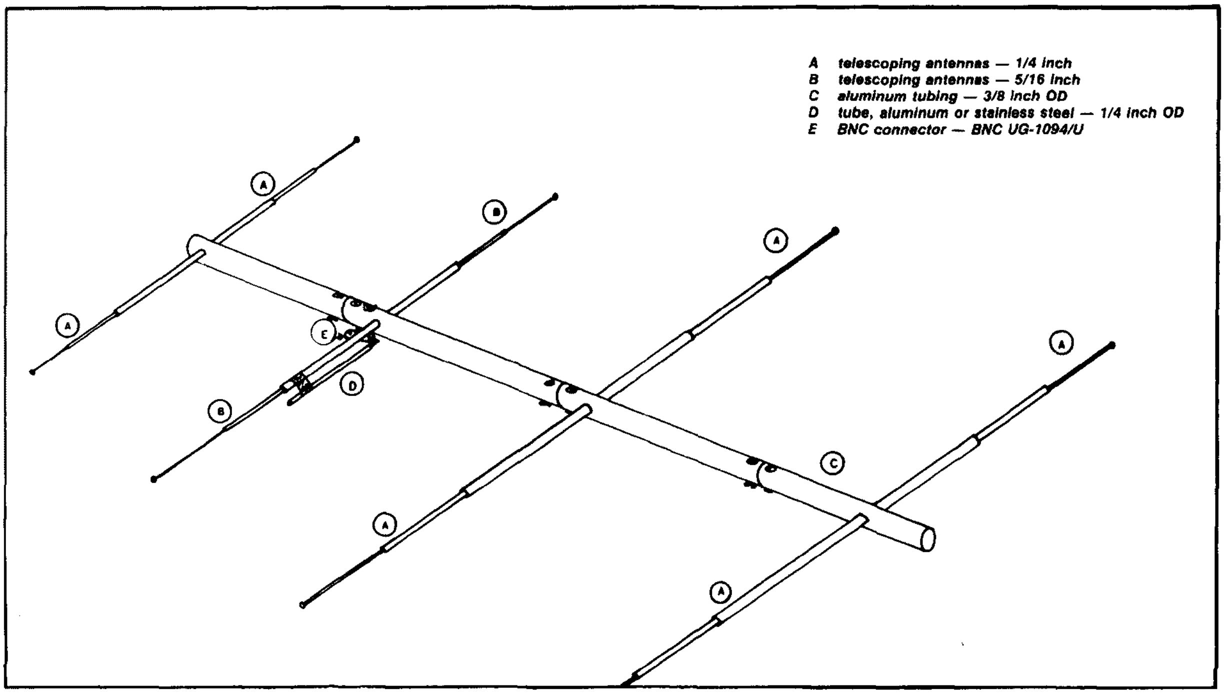


fig. 1A. Fully assembled 4-element 2-meter beam.

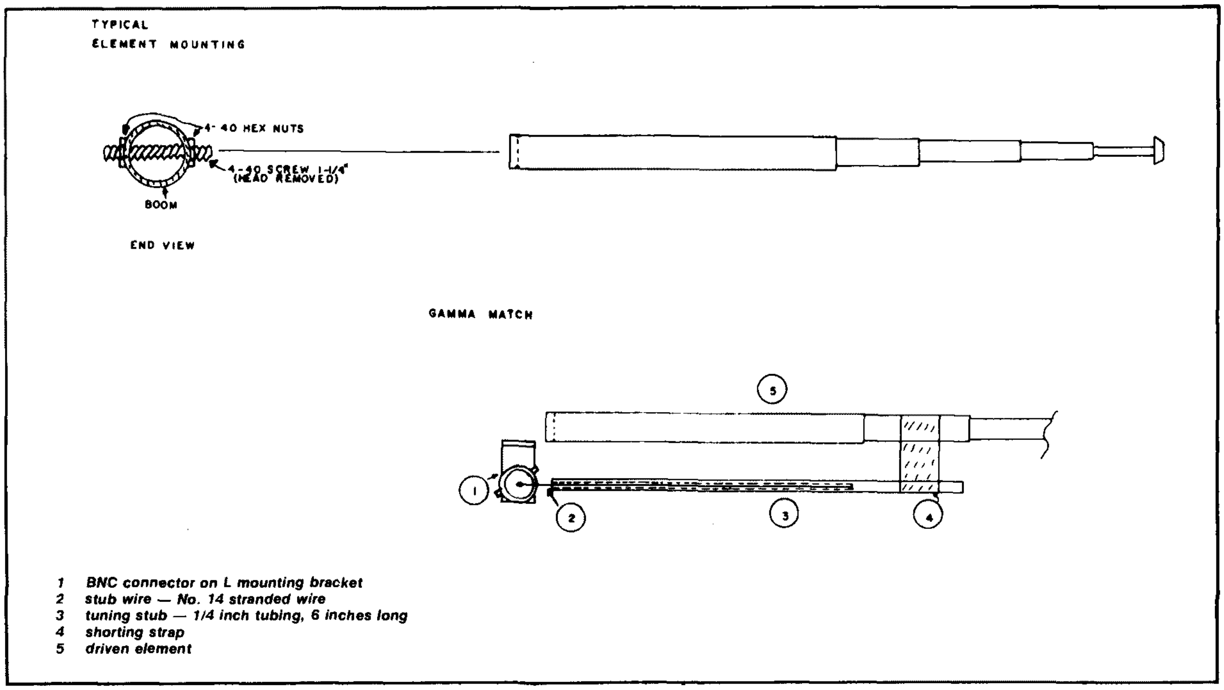


fig. 1B. Element mounting and gamma match details.

The bracket supporting the BNC connector should then be placed under the boom slightly behind the mounting screw for the driven elements. Drill a 4-40 clearance hole through the boom and the bracket. A 4-40 × 3/4 inch (1.9 cm) screw and washer with wing nut is used to allow disassembly to a small size.

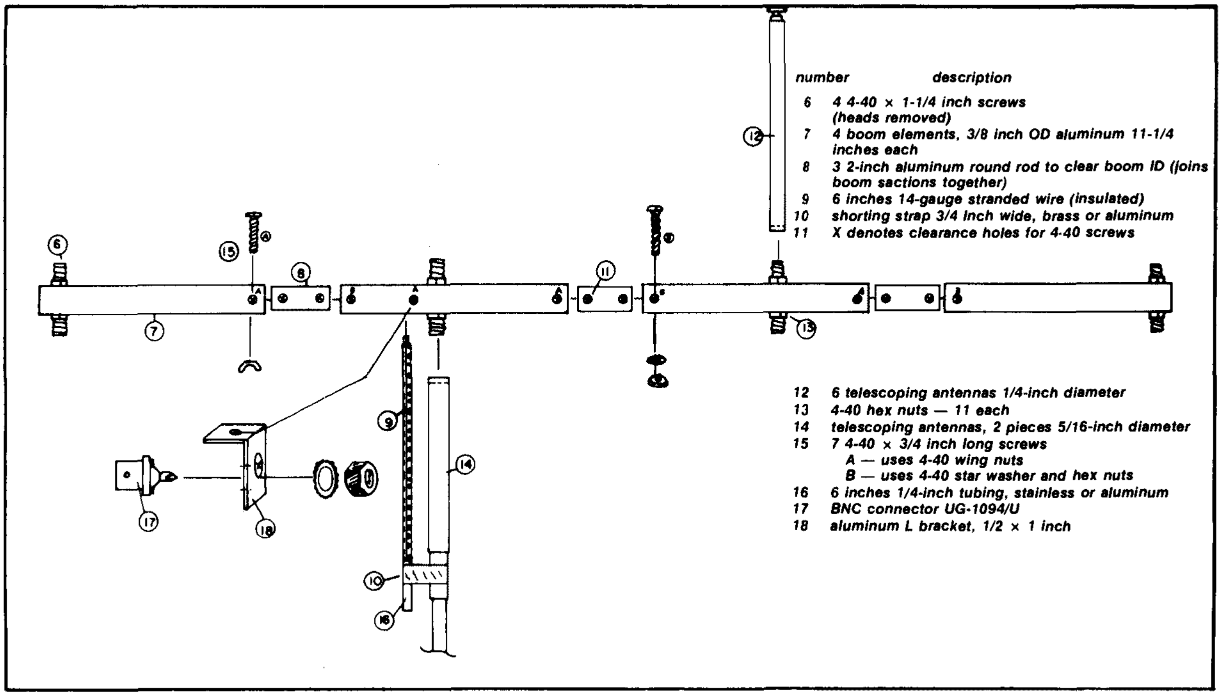


fig. 1C. Assembly details illustrates simplicity in construction and use.

Parts list.

telescoping antennas:

reflector-director (0.24″ diam.)	6 Mouser Electronics No. 43AR106, $1.19
driven (0.28″ diam.)	2 Mouser Electronics No. 43AR105, $1.94

Available from Mouser Electronics, 11433 Woodside Avenue, Santee, California 92071 ($5 service charge applied to orders under $20.)

Normally the whole gamma match is taken apart along with the one driven element to which it is attached for ease in reassembly.

Construction of the antenna is now complete. Assembly for testing and use will require eight telescoping whip antennas (see parts list). Re-tap the holes for the 4-40 screws in the ends to allow mounting on the boom.

tuning

To tune the antenna, assemble the boom, the elements, and the gamma match. Pull the director and reflector out to maximum length. Set the driven elements 19 inches (48.26 cm) apart. Then tune with the gamma match by sliding the shorting stub in and out and the 6-inch (15.24 cm) tubing over the No. 14 wire and adjusting the length of the driven elements. Once the SWR is determined for the frequency you plan to use, secure the gamma in place and measure the length of the driven elements. To simplify setting the length of the driven elements each time, paint a scale on the bottom of the boom to allow accurate positioning each time.

Storage is simple. Unscrew the elements from the boom. Take the gamma match and BNC bracket off. Unscrew the three wing nuts on the boom and remove the screws. Collapse the elements and slide them into the boom. Reinsert the screws, add the wing nuts, and a compact four-element beam is yours. Gain is about 4 dB. From here to Tulsa and other surrounding cities, it's possible to bring up repeaters on 1 to 2 watts of power.

Additional elements can be added by extending the boom to allow each element to be 14-1/2 inches (36.83 cm) from the last. Use your imagination as to ways to mount it for your own use.

Note: this idea was submitted to my employer, Phillips Petroleum Company, under a patent agreement. After lengthy review, it was released to me for the purpose of making it available for use by fellow Amateurs. Commercial use of this design is not permitted under the terms of the release letter signed by Phillips and dated June 2, 1982.

ham radio

the ham notebook

increased undistorted TS-930S headset audio

Many operators prefer headset operation employing a relatively high audio output level combined with low RF gain. This mode tends to preclude AGC capture from strong adjacent signals and tends to minimize noise. Using the very popular Kenwood TS-930S in this manner of operation, peak audio distortion is evident depending upon the particular headset, the signal levels, and the gain settings employed.

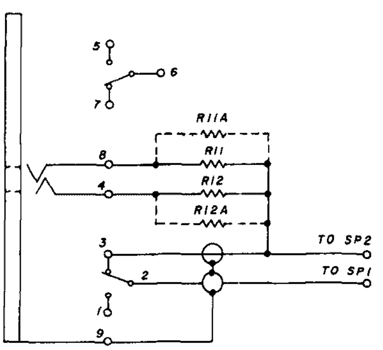


fig. 1. TS-930S headset jack circuit.

A simple solution to this condition is to decrease the value of the 1/2-watt, 100-ohm voltage divider resistors, R11 and R12, connected to the headset jack on the front panel as depicted in **fig. 1**. It is necessary to remove the triangular corner brace plate adjacent to the jack to gain access to the resistors. The plate is held in place by five easily removable screws. In my project, I paralleled each of the existing resistors, R11 and R12, with a 110-ohm 1/4-watt resistor, which resulted in at least 6 dB of additional audio without any noticeable distortion, even at uncomfortable levels.

An additional advantage of this modification for operators wishing to correct dissimilar auditory response is that by proper selection of the new, *paralleling resistors, the dissimilar* response can be easily corrected. This, however, will be possible only for those using stereo headsets, since the jack circuit was designed to accommodate both stereo and monaural headsets for a monaural output.

Using the external speaker jack on the rear panel was a first attempt to solve this problem. Although this approach provided audio, it revealed an unacceptable underlying white noise while using CW selectivity. External padding could be used to accomplish the same result as I obtained internally, but I prefer a "clean" fix, without any external connections, and the convenience of the continued use of the front panel jackset jack.

Marv Gonsior, W6FR

TR-2500/2600 2-channel programming

The instruction manuals for the Kenwood TR-2500 and TR-2600 describe how to program ten channels into the memory — but what if you wanted to listen to only two? I've found out that this feature comes in handy when I am monitoring the local emergency nets and don't want to scan any additional frequencies for fear the scan cycle will stop on something in which I have no interest. This is how it can be done following the procedure in the PROGRAM (BAND) SCAN:

Set the lower frequency you wish to receive.

Example: 1. Enter 6.640
2. Press "F" AND "#"

Set the highest frequency you wish to receive.

Example: 1. Enter 6.730
2. Press "F" AND "#"

Now repeat the second entry of the highest frequency.

Example: 1. Enter 6.730
2. Press "F" and "#"
3. Listen for the BEEP
4. Press "F" and "★"

Joel Eschmann, K9MLD

product
REVIEWS

meet Dr. QSO — on-the-air simulator — new from AEA

How many new licensees do you know who've never been on the air because of a real bad case of "key fright?" I can think of quite a few. Until now, I've never been certain of just what to tell them to encourage them to get them on the air.

AEA's new C-64 computer program, Dr. QSO, is a Morse code trainer that lets you simulate everyday Amateur communications without actually having a transmitter. When AEA's Dr. DX came out, I recall being utterly amazed at the realism of the unit — how it simulated, to practically the last detail, almost every aspect of on-the-air communications. The new Dr. QSO has that same quality. It should be a real boon to new licensees and prove to be a valuable training aid in Novice and upgrade classes.

All stations that you hear are generated randomly by the computer using currently issued call signs. At the lower end of the bands (the Extra sub-bands) you'll find code speeds similar to what you'd hear on any given night. As you move out of the Extra portion of the bands, you'll find that code speeds slow down until you get to the Novice band, where speeds range from about five to ten words per minute.

Conversations are very similar to those you would hear on the air. Signal reports and names are given, and equipment, the weather, locations, and antennas are all "discussed." Should you miss a piece of information, you can ask the computer for a repeat — and get one! You can also ask the station to slow down or speed up — and it does.

installation

Whenever you're working with a computer, it's a good idea to turn everything off before you insert a new program card into the machine, to eliminate the possibility of damage should there be any static charges. AEA recommends doing this with Dr. QSO, and it's good advice. The program card simply slips into the back of the computer, label side up.

Next, you plug a Morse code key into the back of the unit. You can use a straight key, bug or electronic keyer; electronic keyers are preferred. (Computer decoders sometimes have trouble deciphering a fist because of irregular characters and spacing.)

If you don't have a keyer, or just happen to prefer keyboard operation, press the British "pound" symbol. You can now communicate with Dr. QSO via keyboard.

program set-up

When the program is turned on, you'll see a transceiver at the top of the screen showing the frequency status of both the transmitter and receiver and level indicators for the volume and bandpass filter. You quickly realize that Dr. QSO comes very close to simulating a real transceiver.

Your next step is to load the text file and pick your operating aids. You select from either the cartridge (by pressing 0) or from custom messages on a disk (by pressing 1-9). Choose your code speed, set the volume, adjust the color and contrast levels on the monitor, and you're all set to go.

operation

A typical QSO looks like this:

CQ CQ CQ CQ CQ DE KB2NYA KB2NYA KB2NYA
KB2NYA DE KB2BOF KB2BOF <AR>
KB2BOF DE KB2NYA GE OM TNX FER CALL – UR RST 57N QTH IS OSSINING NY? OSSINING NY ES NAME IS CURLY CURLY – SO HOW COPY? <AR> KB2BOF DE KB2NYA K
. . . and so on.

Although the program seems to be capable of generating innumerable QSOs of virtually unlimited variety, it's also possible to write your own script. If you have a Commodore disk drive and elementary programming skills, you can write original QSOs and even trade with friends to make the Dr. QSO experience even more realistic. The user manual that accompanies Dr. QSO provides more than adequate instructions.

problem solving

With AEA's thorough documentation and high-quality parts selection and workmanship, there's no reason, really, to anticipate problems with Dr. QSO. Still, for questions or problems that may arise, AEA supports the program with a user hotline available during normal business hours (Pacific time, since AEA's headquartered on the west coast). Supplying your "software support number," furnished with every user's manual, gets you patient, knowledgeable assistance without delay. A parts list and diagram of the board are included in the manual so you can do your own troubleshooting, should the need ever arise.

questions

If you ever have the chance to sit down with this program at a Ham show, you may be tempted to "Ask Dr. QSO" for more information after just a few minutes of "communicating" with the imaginary Hams "created" for your operating pleasure and practice. Don't bother . . . Dr. QSO is, after all, just the software brainchild of some clever programmers at AEA. For information, you're better off writing to AEA at P.O. Box C2160, Lynnwood, Washington 98036-0918.

On the other hand, you might just try asking Dr. QSO . . .

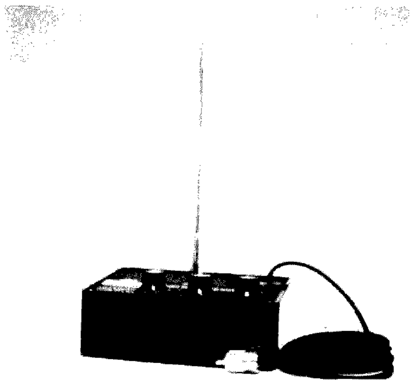
N1ACH and KA1LBO

Circle #301 on Reader Service Card.

MFJ 1621 portable antenna

Would you like to operate portable, but don't want the hassle of installing an antenna?

You won't have any more excuses if you have MFJ's new portable 40-10 meter antenna. Perfect for vacation trips — even DX'peditions — the small lightweight unit "telescopes" for easy stowing in a travel bag. The antenna also comes with 50 feet of coax so you can locate it *after* tune-up to minimize RF interference to your radio or other sensitive electronic equipment. The unit also includes a field strength meter to simplify tune-up.



general information

The MFJ portable antenna is really such a neat idea that I'm surprised it took so long for someone to market it. The unit uses a radiating element that's simply an expandable replacement antenna for auto or shortwave radios. The antenna is tuned by a series capacitor and parallel inductor to ground in a simple "L" configuration. The inductor switch is clearly marked by band with the approximate inductance that will be needed to tune the collapsible whip antenna. The tunable capacitor is also marked so you can establish "pre-sets" for easy tune-up.

MFJ recommends that you encase the antenna in plastic tubing to prevent electric shock if someone should inadvertently touch the antenna while transmitting.

specifications — MFJ 1621

telescoping whip	54 inches
box	5-1/2 × 6-3/4 × 2-1/4 phenolic
RG-58	50 feet
weight	approximately 2 pounds
power	300 watts PEP
bands	40-10 meters

installation

To use MFJ's portable antenna, all you need to do is attach the PL-259 to the back of your transceiver, set the inductor to the appropriate band, set the capacitor to mid-range and apply low power. Varying the capacitor should drop

the SWR to less than 2:1. If it doesn't, change the inductor setting and try again. If more than one setting gives you a low SWR, use the one that gives you maximum power output.

Once this is accomplished, move the antenna away from your operating position — you're almost ready to start working DX. A final check to make sure your SWR is within limits is all that's required before you turn up the power (300 watts PEP maximum).

use

Setting up and tuning this antenna takes just a few seconds and is really quite simple. While no DX was worked during the test period — the bands were in very poor shape — signal reports were more than adequate for the kind of antenna being used. Potential users should not expect this antenna to perform as well as a full signal beam or dipole; it will, however, get you on the air with a more-than-acceptable signal.

Priced at $79.95, the MFJ 1621 will give years of excellent performance and fun. Contact MFJ Enterprises, Inc., Box 494, Mississippi State, Mississippi 39762, for more information.

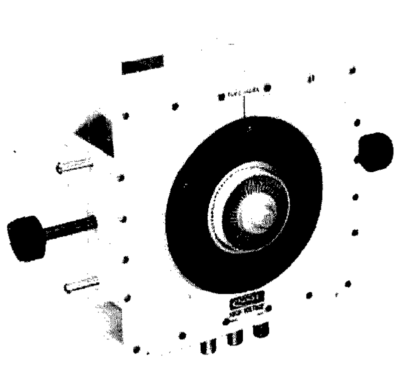
N1ACH

Circle #302 on Reader Service Card.

NEW products

compact VHF power tubes

Varian EIMAC has unveiled four new VHF power tubes designed for power amplifier applications. Developed by Varian EIMAC's Salt Lake Division, the tubes provide improved performance in a compact design. The tubes include liquid-cooled models that use new highly effi-



NEW products

cient heat exchangers to significantly reduce the coolant flow requirements.

Of special interest to Amateurs is the EIMAC 4CX1500BC, a ceramic-to-metal, high-gain power tetrode for service in VHF TV and RF linear power amplifiers. Forced-air cooled with an anode dissipation of 1500 watts, its sophisticated internal structure allows it to operate at full ratings to 450 MHz.

For additional information or literature, contact Varian EIMAC, 1678 Pioneer Road, Salt Lake City, Utah 84104.

Circle #303 on Reader Service Card.

dual-band mobile elements

Larsen has launched a series of dual-band antennas for dual-band Amateur radios, while maintaining its usual high performance standards for both 2 meter and 70 centimeter Amateur bands. The new design incorporates a half-wave element for two-meter (144-148 MHz) Amateur band and collinear elements for 70 cm (440-450 MHz) Amateur band. One antenna conveniently serves both bands while delivering exceptionally high performance. The self-resonant design needs no ground plane, allowing mast installation on boats and base stations with standard Larsen BSA-K hardware.

For more information, contact Larsen Electronics, P.O. Box 1799, Vancouver, Washington 98668.

Circle #304 on Reader Service Card.

shared repeater tone panel

Communications Specialists has announced the introduction of its new TP-38 Shared Repeater Tone Panel. Microprocessor controlled, the TP-38 provides all 38 EIA standard CTCSS tones to allow up to 38 subscribers without the need to purchase additional cards or programming. All features are user-programmable and provided with each unit at one price. Built-in time and hit counters record the activity of all CTCSS tones on the repeater channel. The TP-38 has an ultra low current drain for solar or battery powered repeater sites, and is static and lightning protected. A non-volatile memory retains programming if a power loss occurs. A LED display (which may be turned off to conserve power) shows all received CTCSS tones when they occur, whether they are active in the panel or not. An automatic self-test is activated with each power-up. Any of the 38 available tones may be initiated from the repeater to call down a user for testing purposes. With the addition of the TP-DTMF, an optional, low cost DTMF module, all functions may be performed remotely, using a common 12 or 16-button touchtone pad (secured with a five-digit security code). This allows any of the 38 tones to be remotely turned on or off, and allows the time and hit information to be interrogated remotely.

The TP-38 Shared Repeater Tone Panel and the TP-DTMF Remote Control Module are both available for immediate delivery from stock and are covered by a full one-year warranty. Prices are $595.00 for the TP-38 and $59.95 for the TP-DTMF.

A catalog and further information are available from Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665-4296.

Circle #305 on Reader Service Card.

compact antenna

The Isotron 160 from Bilal Company is only 22 inches tall, 16 inches wide, and 15 inches deep. Although this may seem small, the Isotron 160 has a total surface area of over 900 square inches, 200 square inches more than a 1/2-wave dipole made from No. 12 wire. It has a tested bandwidth of 100 kHz within a 2:1 SWR and will handle the full legal limit of power. It is also adjustable anywhere on the 160-meter band.

Like Bilal's other models, the Isotron 160 does not require any tuning devices or radial system. The hardware is stainless steel and plated and comes complete except for the mast.

The price of the Isotron 160 is $149.95, plus $5.50 for shipping.

For further details, contact Bilal Company, S.R. 2, Box 62, Eucha, Oklahoma 74342.

Circle #306 on Reader Service Card.

HF slopers

Sultronics Amateur Radio announces the introduction of the "second generation" of its compact HF sloper antennas for the 160-80-40 meter bands. Featured are two models: the SS-2A Duoband Sloper which covers 80 and 40 meters, and is 45 feet long and the SS-3A Triband Sloper, which covers the 160-80-40 meter bands and is 60 feet long.

Both models feature standard 50-ohm coaxial feed and "no-trap" construction. Second generatin slopers use only stainless steeel hardware and have heavier duty No. 12 solid copper drawn wire for more strength and bandwidth, an Amphenol coax connector, and a heavy duty aluminum tower mounting bracket. Both models are easily tuned for resonance by following the illustrated instruction manual.

The SS-2A Duoband (80-40) Sloper is priced at $27.95, while the SS-3A Triband)160-80-40) Sloper is priced at $39.95 (ppd).

For more information, contact Sultronics Amateur Radio, 1587 U.S. 68 North, Xenia, Ohio 45385.

Circle #307 on Reader Service Card.

flea market

RATES Noncommercial ads 10¢ per word; commercial ads 60¢ per word **both payable in advance**. No cash discounts or agency commissions allowed.

HAMFESTS Sponsored by non-profit organizations receive one free Flea Market ad (subject to our editing) on a space available basis only. Repeat insertions of hamfest ads pay the non-commercial rate.

COPY No special layout or arrangements available. Material should be typewritten or clearly printed (**not** all capitals) and must include full name and address. We reserve the right to reject unsuitable copy. **Ham Radio** cannot check each advertiser and thus cannot be held responsible for claims made. Liability for correctness of material limited to corrected ad in next available issue.

DEADLINE 15th of second preceding month.

SEND MATERIAL TO: Flea Market, Ham Radio, Greenville, N. H. 03048.

ESTATE SALE for the family of N5CIP. All equipment in mint condition, some brand new. TR7, PS7, R7, MN2000, MN2700, WH7, P75, SP75, HAL 3100ASR, ST6000, AEA 150, IC271H, SPS20, TB5EM and other parts. Send SASE for complete list. Call KD5E (409) 632-5625 or (409) 634-4317.

FREE 3 month subscription. Send callsign to Westlink Report-H10, 28221 Stanley Court, Canyon Country, CA 91351. New only!

THE GOOD SAM HAMS invite RV operators to check in the Good Sam Ham net 14.240± Sundays 1900Z also 3.880± Tuesdays at 2359Z. Net control N5BDN, Clarksville, IN.

SCHEMATIC: Radio receivers 1920's/60's. Send brand name, model No., SASE. Scaramella, PO Box 1, Woonsocket, RI 02895-0001.

LET THE WORLD KNOW what you would really rather be doing! Get your "I'd rather be DXing" bumper sticker with "Amateur Radio" written down the sides. Red and black, vinyl and removable, so it won't hurt your car. $2.00 plus 50¢ postage to Bumper Sticker, PO Box 6349, Evansville, IN 47712.

PHELPS DODGE duplexers for two meters, model No. 497, 6 cavity, $600 each. HAL ID-1A repeater ID'er, 19" rack mount, 110V power supply, $50. VHF Engineering two meter repeater complete with logic, $325. K1KYI, Richard Fairweather, Harrisville, RI 02830. (401) 568-3468 between 8-9:30 PM EDT.

2 METER AMP KIT: 8877 legal limit kit $395. 3CX800A7 900W kit $325. Also HV power supplies, CX600N relays, parts and EME newsletter. SASE for catalog. 2 Meter EME Bulletin, 417 Staudaher St., Bozeman, MT 59715

BACK ISSUES QST, 73, Ham Radio. Stamp for list. Koepke, 6 Katherine Rd., Albany, NY 12205.

WANTED: Regency XL2000 transceiver or new 10 channel 148-162 MHz or Wilson WH2510. PO Box 929, Blacksburg, VA 24060-0929. (703) 382-4458.

WANTED: Ten-Tec Argosy 525. Mike Kaufman, K6VCI, 107 Suffield Avenue, San Anselmo, CA 94960.

CABLE TV CONVERTERS DESCRAMBLERS. Jerrold, Hamlin, Zenith — many others. Factory units/lowest dealer prices. Complete illustrated catalog, $2.00. Pacific Cable Co., Inc., 7325-1/2 Reseda Blvd., Dept. 1005, Reseda, CA 91335. 818/716-5914.

NEW VLF 2 kHz-500 kHz products by K1RGO. L-101B VLF converter, L-201 VLF broadband preamplifier, L-400 VLF active antenna and more. Write for free catalog. LF Engineering Co., 17 Jeffry Road, East Haven, CT 06512.

ICOM AND KENWOOD SEPARATE NEWSLETTERS: 5 years of back issues! Now available TS930S & 430S users modifications supplement. Send 39¢ SASE for 12 page FREE Brochure to: International Radio, Inc., 1532 SE Village Green Dr., Suite L, Port St. Lucie, FL 33452.

HELP: Need schematic for Knight 2-meter transceiver Model TR-108. Will pay for copy. Ph. (209) 626-4219. KB6JDT, Donald M. Cox, 318 Park Blvd., Orange Cove, CA 93646.

SIGNAL GENERATORS, URM-25D, 10 kHz thru 50 MHz $245.00, URM-26B, 4 MHz thru 405 MHz $245.00, HP614A, 900 MHz thru 2100 MHz $345.00, HP618B, 3.8 GHz thru 7.6 GHz $375.00, HP608C 10 MHz thru 480 MHz $345.00, TS-510/U 10 MHz to 420 MHz $295.00. All lab calibrated. Have good stock so order today. We accept M/C, VISA, or check. FOB Otto. Immediate shipment. Phone Bill Slep 704-524-7519. Slep Electronics Company, Highway 441, Otto, NC 28763.

CABLE TV CONVERTERS & EQUIPMENT: Plans and parts. Build or buy. FREE information. C&D Electronics, 626 Flowerdale, Dept. HR, Ferndale, MI 48220.

TR-7 USERS: NB-7 Noise blanker (new) $65.00 ppd. HS-75 headset (padded) new $14.00 ppd. SL-300, SL-500 CW filter $55.00 ea. ppd. K3UKW, Tony Musero (215) 271-8898.

FOR SALE: Kenwood TR-7800 two meter transceiver $200.00/offer; Regency & Halicrafters 150-174 MHz receivers $15.00 each; Realistic SSB/AM CB $75.00. WB9YBM, Klaus Spies, 8502 Oketo, Niles, Illinois 60648.

ELECTRON TUBES — Radio & TV types 75% off list price — huge inventory! Also industrial types. Send for Free Catalog today or call Toll Free (800) 221-5802. Box HC, Transleteronics, Inc., 1365 — 39th Street, Brooklyn, NY 11218. See our display ad this issue.

IMRA, International Mission Radio Association helps missionaries. Equipment loaned. Weekday net, 14.280 MHz, 2-3 PM Eastern. Eight hundred Amateurs in 40 countries. Brother Frey, 1 Pryer Manor Road, Larchmont, New York 10538.

ELECTRON TUBES: Receiving, transmitting, microwave . . . all types available. Large stock. Next day delivery, most cases. Daily Electronics, PO Box 5029, Compton, CA 09224 (213) 774-1255.

CHASSIS and cabinet kits. SASE K3IWK.

CUSTOM MADE embroidered patches. Any size, shape, colors. Five patch minimum. Free sample, prices and ordering information. Hein Specialties, Inc., Dept. 301, 4202 N. Drake, Chicago, IL 60618.

CX7 REPAIRS. 415-549-9210.

RTTY-EXCLUSIVELY for the Amateur Teleprinter. One year $7.00. Beginners RTTY Handbook $8.00 includes journal index. PO Box RY, Cardiff, CA 92007.

TRS-80 Model I/III/IV owners. HF antenna design program calculated dimensions for dipole, yagi, and quad antennas. $14.95 (cassette) + $2.00 s&h to Cynwyn, Dept. H, 4791 Broadway, Suite 2F, New York, NY 10034.

OLD RADIO transcription discs wanted. Any size, speed. W7FIZ, Box 724 HR, Redmond, WA 98073-0724.

RUBBER STAMPS: 3 lines $4.50 PPD. Send check or MO to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

RF DESIGN PROGRAM calculates stability, gain, noise figure circles, and more. For C-64 and IBM-PC, $29.95. Jim Allyn, N7JA, 7000 Litchford Road, #26, Raleigh, NC 27609.

AMATEUR RADIO CLASSES: Chelsea Civil Defense will sponsor evening classes at Chelsea High School starting October 16, 1985, for Novice (basic level ham license), and Tech/General licenses. Minimal cost for materials only. For information: Frank Masucci, 136 Grove Street, Chelsea, MA 02150.

Coming Events

ACTIVITIES

"Places to go..."

INDIANA: The Allen County Amateur Radio Technical Society's 13th annual Fort Wayne Hamfest, Sunday, November 10, 8 AM to 4 PM, Allen County Memorial Coliseum, Coliseum Blvd., U.S. 30. Indoor tables available $8.00, AC power extra. Premium tables with AC $20.00 each. Admission $3.50 advance, $4.00/door. Children under 11 free. Ladies' activities, forums, banquet Saturday night. Nearby motels and restaurants. VE exams Saturday, November 9 advance registration only. Talk in on 146.28/88. For information or reservations: AC-ARTS Hamfest, PO Box 10342, Fort Wayne, IN 46851.

GEORGIA: Ham Radio and Computer Expo '85, sponsored by the Alford Memorial Radio Club of Stone Mountain, Gwinnett County Fairgrounds, Lawrenceville, 20 minutes NE of Atlanta. Saturday November 2, 9-5; Sunday, November 3, 9-4. Admission $4.00 advance, $5.00/door. FCC license exams both days. Free cookout Saturday night. Activities for the whole family. RV sites with hookups on site. Talk in on 146.16/146/76. For information and reservations: Alford Memorial Radio Club, PO Box 1282, Stone Mountain, GA 30086. (404) 476-2944.

NEW JERSEY: Stateline Hamfest sponsored by the Stateline Radio Club, NY and NJ, November 16, St. Andrews School, 120 Washington Avenue, Westwood, NJ. Doors open 8 AM. Tickets $3.00/gate. Tailgaters $6.00/space. Vendors $10.00 advance by 10/31/85. $13.00/gate. VEC testing. Food and refreshments. Talk in on 146.835 K3LSA Rpt. For information: Stateline Hamfest, Stateline Radio Club, PO Box 325, Montvale, NJ 07656 or call Fred, N2ATI (201) 664-5320.

ILLINOIS: EARS, Inc. Hamfest, Sunday, November 10, Harlem Community Center, 900 Roosevelt Road, Machesney Park. Advance tickets $3.00/SASE. $4.00/door. Inside tables $5.00 each. Talk in on 146.01/.61. For information: EARS, Inc., PO Box 4291, Rockford, Il 61110.

IOWA: The Muscatine and Iowa City Amateur Radio Clubs will co-sponsor the SE Iowa Hamfest, Sunday, October 6, rain or shine, West Liberty Fairgrounds, West Liberty. Gates open 7 AM. Indoor/outdoor flea market. Saturday night camping available on grounds, nominal charge. ARRL/VEC exams, preregistration suggested. Tickets $3.00/advance and $4.00/gate. For further information KEOY, Tom Kramer, 905 Leroy Street, Muscatine, IA 52761. (319) 264-3259.

OHIO: Marion Amateur Radio Club's 11th annual "Heart of Ohio" Ham Fiesta, Sunday, October 27, 0800 to 1600 hours, Marion County Fairgrounds Coliseum. Tickets $3.00/advance. $4.00/door. Tables $5.00. Check in on 146.52 or 147.90/30. For information, tickets or tables: Ed Margraff, KD8OC, 1989 Weiss Avenue, Marion, Ohio 43302. (614) 382-2608.

ALABAMA: S.P.A.R.C. 1st annual Swapmeet and Packet exhibit, Lee County Fairgrounds, US 431, October 19, Opelika/Auburn. 10 AM to 5 PM. Spaces $5.00 per vehicle advance; $7.00/gate. $1.00 gate donation. Free parking. Packet demonstrations by Bob McGwier, N4HY. Reservations contact: Ray, PO Box 2423, Opelika, AL 36803-2423. For information call Ray (205) 745-2838, Gene (205) 821-8010, Danny (205) 745-7455. Talk in on 147.06/66.

PENNSYLVANIA: RF Hill ARC's 9th annual Hamfest, Sunday, October 29, Pennsylvania National Guard Armory, Rt. 152, Sellersville. Indoor ($8) and tailgating ($6) space available. Sellers set 6 AM. Buyers admitted 8 AM. Tickets $3.00. Nonham spouses and children free. For indoor spaces call WB3AIG (215) 874-4800, ext. 515. Tailgating space first come, first served. Talk in on 144.71/145.31, 146.28/88 and 146.52.

TENNESSEE: The 7th annual Amateur Radio and Computer Convention, October 26 and 27, new Convention and Trade Center, Chattanooga. Free admission. All indoors. VE exams both days. 8' flea market tables $6/day; $10/both days. Power $5 extra. Talk in on 146.19/.79. For information: Hamfest Chattanooga, PO Box 3377, Chattanooga, TN 37404 or call Nita Morgan, N4DON (404) 820-2065.

FLORIDA: The 10th annual South Florida ARRL Suncoast Convention, October 12 and 13, National Guard Armory, St. Petersburg. Note new location: QCWA luncheon Saturday. Luau Saturday night and Ladies' luncheon Sunday. Special demonstration on packet radio for beginners and advanced operators. VE exams Saturday morning. Registration $3.00. $4.00/door. For tickets and hotel rooms: FGCARC, 1556-56th Avenue N, St. Petersburg, FL 33703. Checks payable to FGCARC.

MINNESOTA: Hamfest Minnesota and computer Expo, sponsored by the Twin City FM Club, November 2, Richfield High School, 7001 Harriet Avenue S, 8 AM to 3 PM. Admission

$3.00/advance; $4.00/door. FCC exams, giant indoor flea market. Talk in on 16/76. For information: Clyde R. Green, N0DVP, 5406 Zealand Avenue N., New Hope, MN 55428 or Twin City FM Club, Box 555, Minneapolis, MN 55440.

MASSACHUSETTS: Giant electronics auction sponsored by the Hampden County Radio Association, November 1, Granger School, Feeding Hills starting 7:30 PM. Sponsor takes 10% commission on all sales. Proceeds benefit club projects including building an emergency communications system. No admission charge. Public invited. Refreshments available. For information: Ron Beauchemin (413) 739-5228.

TEXAS: The Austin Amateur Radio Club's Fall Swapfest, October 19, 8 AM to 1 PM, Manchaca Fire Hall, Manchaca, Farm Road 1626 west of IH-35. Covered and open-air spaces. Sellers may bring own tables. Indoor tables $2.00 each, first come, first served. Dealers welcome. Admission is free. Talk in on 146.19/.79 or 16.34/.94. For information: Jim Strohm, KA5UXC, 1743 Cricket Hollow Drive, Austin, TX 78758. (512) 837-5423 or 837-4352. SASE for map and information.

TENNESSEE: The 5th annual Tri-Cities Hamfest sponsored by the Johnson City and Kingsport Amateur Radio Clubs, October 19, Appalachian Fairgrounds, Gray. Forums, dealers, flea market and RV hookups. For information: Tri-Cities Hamfest, PO Box 3682 CRS, Johnson City, TN 37601.

MICHIGAN: 3rd annual Hamfest/Electronic Flea Market, October 27, Kalamazoo County Fairgrounds, Kalamazoo. Admission $2.00/advance; $2.50/door. 8' table space $6. Dealer setup 8:30 AM. Doors open 9 to 4. Ham license testing 10 AM, limited walk-ins. For tickets/tables: Ken, KA8RUA, 2825 Lake Street, Kalamazoo, MI 49001.

GEORGIA: Rome Hamfest sponsored by the Coosa Valley ARC, Sunday, October 6, Rome Civic Center, GA Highway 20. Admission free. Homemade Bar-B-Cue and stew. Camper parking available, no hookups. Inside tables $5, outside $3. Fun for the whole family. Talk in on 147.900/300. Contact Buddy Waller, NO4U, 24 Wellington Way, SE Rome, Georgia 30161. (404) 235-5417.

NEW HAMPSHIRE: Hosstraders annual Fall tailgate swapfest, Saturday, October 5, Deerfield Campgrounds. Donation $2 per person sellers included. Profits benefit Shriners' Boston Burns Center. Our May Swapfest gave $6,960. Friday night camping at nominal fee after 4 PM. Talk-in on .52 and 146.40-147.00. For information SASE to Norm, WA1IVB, RFD Box 57, West Baldwin, ME 04091.

OPERATING EVENTS

"Things to do . . ."

October 12: Hermiston Amateur Radio Club's KD7LC will be on the air 1800-0100Z and 1800-2200Z, October 13 to commerate the 104th anniversary of Lewis and Clark's visit to Hat Rock. To receive special certificate send 9x12 SASE to HARC, PO Box 962, Hermiston, OR 97838. November 10: In observance of Veterans' Week, members of the Hamfesters Radio Club, Chicago, will operate from the Robert K. Wade (K9CDH) Memorial Ham Shack, Hines VA Hospital using the Hines Club call K9WFN, 1500Z to 0300Z. 40, 20, 2m FM, 2m USB. For a certificate, send QSL and 9x12 SASE to Hamfesters RadioClub, Inc., Chicago, c/o Robert K. Wade Memorial Ham Shack, Hines VA Hospital, Hines, Illinois 60141.

October 12: Members of the Capeway Radio Club will commemorate the 350th anniversary of Plymouth County (Mass) from October to October 20. For a certificate send a large SASE with 39 cents postage and QSL to: Ray Witt, WA1OWQ, 62 Caldwell St., No. Weymouth, Mass. 02191

October 5: A special event station, KN5D, will operate from the annual International Hot Air Balloon Fiesta, Albuquerque, New Mexico, October 5 through October 13. Also a gateway station through Oscar 10, conditions permitting. QSL to KN5D, PO Box 997, Corrales, NM 87048.

October 27: Laurel ARC will operate special event station W3DQI, 1500Z to 2230Z to help celebrate the anniversary of the restoration of the Montpelier Cultural Arts Center of Montpelier Mansion in historic Laurel, MD. 8x11 certificate for SASE. QSL to LARC, PO Box 91, Annapolis Junction, MD 20701.

October 15: The Colquitt County Ham Radio Society will operate club station WD4KOW from the site of the 8th annual Sunbelt Agricultural Exposition, October 15, 16 and 17. 0900 to 1700 EDST each day. The Sunbelt Expo is at the Spence Field Airbase, near Moultrie, Georgia. A special QSL card is available for those making contact and sending a SASE.

October 19: JOTA — Scouts 28th annual Jamboree on the Air. Look for K2BSA, the BSA headquarters station in Dallas, Texas and *HB9S, the World Scout Headquarters in* Switzerland and other special call signs from many countries. Requests for certificates should go to Jamboree on the Air, 1325 Walnut Hill Lane, Irving, TX 75038 with large SASE. Encourage scouts in Amateur Radio.

October 26: The St. Peters, MO, ARC will operate 1700Z October 26 to 1700Z October 27 from the Daniel Boone Home, Femme Osage Valley, St. Charles County, to commemorate where Boon spent the last two decades of his life. Listen for KB0J. Certificate via Bob Goin, KA0IKU, 3112 Powder Horn Trail, St. Charles, MO 63301.

THE GUERRI REPORT

Ernie Guerri W6MGI

RF effects — the good and the bad

We're all familiar with the convenient way in which microwave ovens quickly cook even large pieces of food. A little closer look at our friendly oven will reveal that considerable effort has gone into making sure that the microwave radiation doesn't leak out, and that the unit won't operate unless the door is closed and locked. There's good reason for this — human tissue is very sensitive to radiation, and the effects are cumulative.

The American National Standards Institute (ANSI) has developed a "whole-body" standard for human exposure to radiation (**fig. 1**).

There is not general agreement on the levels set in the ANSI standard, and these levels have not been adopted by the Environmental Protection Agency. Several countries have standards which are considerably more restrictive than ANSI's, and some permit levels of only 1/10th the U.S. amount. This lack of agreement only serves to emphasize that one should err on the side of conservatism in such a critical area.

It's important to note that the most dangerous frequency ranges include the Amateur bands from about 10 meters through 450 MHz, and that some parts of the body are more sensitive than others. Human eyes, for example, are several times more sensitive to RF than limbs. It would be wise for Amateurs who use handheld transceivers in the VHF/UHF region to make sure that their antennas are pointed well away from their faces when transmitting. Separate antennas are safest for power levels greater than 3 to 5 watts.

Similarly, care should be taken when sending high power to antennas mounted directly on the roof of a house; the field strength inside the house can easily exceed the ANSI limits. *Never* stand directly in front of a high gain VHF/UHF or microwave antenna when applying any appreciable power.

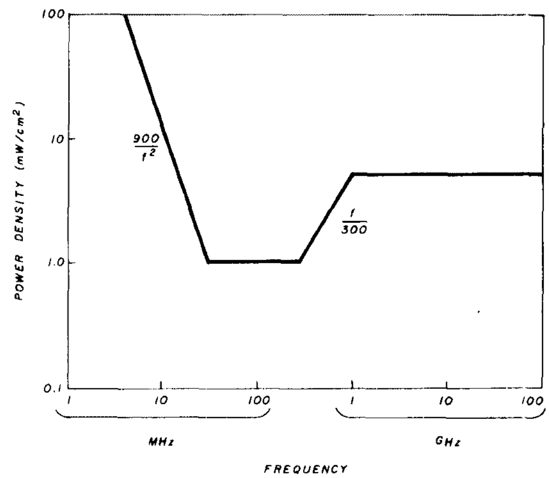


fig. 1. ANSI whole-body RF exposure.

There are positive uses for the application of RF energy to the human body, however. Considerable work is being done to examine the therapeutic effects of various RF fields. The results are encouraging. Both thermal and non-thermal effects have been observed to diminish or eliminate tumors, and pulsed RF can enhance bone growth in the healing of fractures. Experiments in the 30 MHz region have shown that nerves and related tissues can actually regenerate when exposed to levels of about 50 mW/cm² — much higher than the ANSI whole-body standard.

It would appear that RF is like many of the chemicals our body needs. Delivered in just the right quantity, at just the right rate, they make us healthy. Misused, they can be dangerous.

solar cells achieve high efficiencies

We all know that solar cells make fine power sources for calculators and NASA vehicles. But we may soon realize a wider range of benefits from these long-touted devices. Tests at Sandia National Laboratories in New Mexico have confirmed that high temperature gallium arsenide cells can yield efficiencies of over 25 percent. Silicon cells housed in concentrators that focus light on the cell surface have demonstrated conversion efficiencies of 17 percent, with prospects for improvement of another 1 to 2 percent in the near future. The cost of the overall cell-lens assemblies is still high — nearly $10/watt in small quantities. However, with such significant improvements in efficiency, researchers can soon concentrate on ways to

reduce costs and contemplate true commercialization of solar electric panels.

right on

The stirring vision of a pilot in the cockpit with a silk scarf, goggles, and only a compass to guide him, long ago gave way to an image of a much calmer environment, complete with autopilot. Then about 20 years ago the autopilot began to be replaced — on long flights — with inertial navigation systems (INS) offering typical errors of less than one degree/hour of drift.

Continued improvements have brought us to the point of laser beams illuminating reflective spheres only a few microns in diameter, as the main inertial element in a gyro. These little spheres spin at more than 200,000 RPM, which means that they don't drift much with time. Complex electronics keep track of the exact position of the sphere and correct the overall gyro accuracy by a position fix against a GPS (Global Positioning Satellite). The resulting navigation system permits flying intercontinental distances with only a few hundred meters of total course error. To top it off, laser gyros on a chip are expected to be available in 1988-89.

Now, imagine a modern Robin Hood with his sheaf of "smart" arrows — laser guided and microcomputer controlled, each with a memory for its specific target: deer, fox, rabbit . . . and so on. He wouldn't even have to *shoot* the arrows — just throw 'em — by the handful! What hath technology wrought?

ham radio

Invitation to Authors

ham radio welcomes manuscripts from readers. If you have an idea for an article you'd like to have considered for publication, send for a free copy of the *ham radio* Author's Guide. Address your request to *ham radio*, Greenville, New Hampshire 03048 (SASE appreciated).

✓ 204

NOVEMBER 1985 / $2.50

annual **RECEIVER** issue

ham radio magazine

build this compact 75-meter monoband transceiver

hr

focus on communications technology

stable LO's for microwave receivers • 10-80 meter homebrew receiver • adding digital frequency readouts to transceivers • scanner for CB-to-10 meter conversions • external receiver product detector • build a handy RF probe • plus W6MGI, W1JR, W6SAI, W9JUV, and K0RYW

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NOVEMBER 1985

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contents

REFLECTIONS

muchas gracias

Yesterday, September 20, a massive earthquake struck Mexico City. Early reports from the networks and wire services told of massive damage in Mexico City, with 25 per cent of the downtown area totally destroyed, and thousands of deaths and injuries. Ships have been reported missing and a tsunami wave has severely damaged part of the Mexican coastline. All modes of communications — save one — are down and authorities have no idea of when service might be restored.

The one mode that's still functioning is Amateur Radio. I've spent quite a while listening to the emergency traffic being passed on a number of frequencies; I'm quite impressed with how effectively and efficiently health and welfare traffic is being passed. But there's a personal sense of urgency to my listening this time: my youngest brother is in Mexico City, and my family and I have no way of knowing if he's all right.

Yes, the State Department has a hotline that we can call for information. But the volume of calls received has shut the system down. We've contacted the Red Cross, but they're too busy passing more important medical traffic.

So how do I — or others — use Amateur Radio to find out about the safety of relatives and friends? I'm afraid I really don't know. I usually have the world at my fingertips with just the flip of a switch and the twist of a dial. But right now I feel just about totally helpless in trying to get the information I want.

Propagation into the Mexico City area isn't favorable at this time, and I'm reluctant to jump into a pile-up and add to the confusion. But that's about the only way I'll be able to get a message to the Mexican stations. . . .

There's got to be a better way! Would somebody please tell me what it is?

Craig Clark, N1ACH
Assistant Publisher

Craig's inital feeling of helplessness is completely understandable. The circumstances — a natural disaster that, within 60 seconds, killed thousands, injured many, and caused catastrophic damage to buildings and other structures — is something one doesn't deal with on a daily basis. So it's hard to pull from your experience and function immediately.

Several things are reasonably certain: all conventional power and most telephone service is gone. Few reliable means of communicating with the outside world are available. This is where Radio Amateurs come in — and what a service they provide! Just after the Mexico City earthquake, several stations were on the air. One in particular that I heard was XE1VIC, manned by Vic and Sergio, who did a yeoman's job in passing health and welfare information, medical emergency traffic, and high-level government phone patches.*

With this in mind, what can *you* do to find out about the safety of a relative or friend known to be in a particular area when disaster strikes?

LISTEN CAREFULLY. If there's any activity from the devastated area, someone as competent as XE1VIC is likely to be on the air. If you don't hear such a station, continue listening. Chances are good that other Amateurs, in your country will have information and be discussing it.

IF YOU DON'T HEAR, ASK. Once again, someone will probably know something.

THINK. For the distances involved, time of day, and propagation, one particular band will be optimum. Try it. XE1VIC was 59+ into New Hampshire on 80 meters for hours.

LISTEN FOR INSTRUCTIONS. The control station will make it very clear as to how he wishes to proceed. Many stations will be attempting to contact him. Though it may seem like contest operation at times, remember that a disaster has occurred, and discipline is essential.

BE BRIEF. Once you've gotten the traffic handler's attention, give the information required — nothing more, nothing less. Decide beforehand how you might be able to help.

ACCEPT THE RESPONSE GIVEN. Once he contacted the requested relative or friend, XE1VIC kept his side of the conversation short and sweet. Most of the time the news was very good, and in every case he was extremely reassuring. Remember that many other stations are concerned about their loved ones, too.

OTHERS WILL HELP PASS YOUR MESSAGE. In this case, stations in the Houston area acted in this capacity.

Every emergency, of course, is different. These are just a few suggestions that have been applied in the past and appear to work.

Rich Rosen, K2RR
Editor-in-Chief

Note: Craig's brother returned home safely on September 23.

*XE1VIC was just one of many Mexican stations that should be commended. In addition, the list of U.S. and international stations that contributed to the success of this traffic handling situation in a calm and controlled manner is long, and one we should be quite proud of — proving once again how Radio Amateurs rise to the occasion and provide a real service.

presstop de W9JUV

AMATEUR RADIO'S RESPONSE FOLLOWING MEXICO CITY'S DISASTROUS EARTHQUAKE was, in short, simply overwhelming! Almost before the first major tremors ended, a number of Mexico City stations were on and handling emergency traffic with the U.S. As the hours wore on and the disaster's magnitude became apparent, the activity multiplied—but even after health and welfare input began, and despite a variety of message handling techniques, turnaround was generally quite fast. W9JUV's first two Friday afternoon inquiries came back "We're all OK!" from XE1USA in under five minutes! Amateur Radio's media exposure was also fantastic due to the complete breakdown in commercial communications, which left the Amateur Service as the media's only source of earthquake news for a number of hours.

Media Abuse Of Amateur Radio Was Also Extensive, unfortunately. Not only was there the usual "XE1XX, I have Newt Newshawk of NNUT-TV in the shack to ask you some questions," but all of the networks and news services invaded the Amateur bands for their coordination and logistical efforts! Heard on 20 meters Sunday were equipment requests, personnel assignments, even arguments about overtime and the special menu needs of a network anchor! To its credit, NBC's Mexico City station did take health and welfare traffic and made phone calls when not otherwise occupied; inquiries to the others triggered responses that they were too busy on network business to make casual contacts! The FCC is very much interested in this media incursion, and would appreciate receiving specific reports and tapes on it.

One Glaring Weakness Of Amateur Radio Was Quickly Apparent—anxious friends and relatives of persons living in or visiting Mexico City had no idea of how to locate hams willing to relay messages. Though churches, community centers, the Red Cross, and even some Amateur Radio dealers were often able to make referrals, what's badly needed—before another such catastrophe strikes—is a well advertised, accessible conduit through which the public can reach Amateur Radio operators able to communicate with loved ones abroad.

Despite Such Problems, However, Amateur Radio Has Never Served People Better or made a deeper, more favorable impression on the public than it has during Mexico City's agony. "Well done!" to all involved, both those actively participating and the many "stand-bys."

STATE AND LOCAL REGULATIONS MAY NOT PRECLUDE AMATEUR COMMUNICATIONS, the FCC declared in its September 16 decision on PRB-1! Responding to the ARRL's July, 1984, request for a Declaratory Ruling limiting local limitations on Amateur operations and stations, the FCC affirmed "a strong federal interest in promoting Amateur communications" and cited Amateur Radio's value in "providing emergency communications" and "a reservoir of trained operators, technicians and electronic experts," as well as furthering international goodwill. In summary, they said: "State and local regulations that operate to preclude Amateur communications in their communities are in direct conflict with federal objectives and must be preempted."

Though They Did Not Specify Any Minimum Antenna Height, the Commission further stated, "...local regulations which involve placement, screening, or height of antennas based on health, safety, or aesthetic considerations...must be crafted to accommodate reasonably Amateur communications (with) minimum practicable regulation...."

Though The Lack Of A Specified Height Will Cause Some Amateurs Problems, the FCC's action is certainly going to help the vast majority. Also not covered are Amateurs who've signed restrictive covenants, since those are private contracts entered into voluntarily.

Congratulations To The Commissioners And FCC Staff for a courageous decision, since their favorable ruling for Amateur antennas and operations will undoubtedly be cited by other services who'd also like to get out from under the yoke of local regulation.

POSSIBLE FEDERAL SANCTIONS AGAINST "ELECTRONIC EAVESDROPPING" are close to being introduced in the U.S. House of Representatives by Rep. Kastenmeyer. In his "Electronic Communications Privacy Act of 1985," Rep. Kastenmeyer proposes restricting the monitoring of any transmission by "wire, radio, electromagnetic, or photoelectric system..." with exceptions for communications "readily accessible to the public," stations for general public use, distress signals, police or fire, and Amateur or CB stations. Penalties for commercial violators could be $25,000 and a year in prison, others $5,000 and six months.

Though The Goal Is Almost Surely The Protection Of Cellular Radio, the effect on Amateurs and others with a general interest in radio communications could be serious.

APPLICANTS WHO FLUNK AN AMATEUR EXAM NEEDN'T WAIT 30 DAYS to retake the exam, the FCC decided September 16. In its decision to eliminate the waiting period entirely, the Commission agreed a delay had little benefit and dropping it could permit applicants who failed the first day of a two-day hamfest to review problem areas for a second-day try.

DPØSL WILL BE THE CALLSIGN FOR THE UPCOMING EUROPEAN STAFFED SPACEFLIGHT, now set for launch about October 30th on Space Shuttle Flight 61-A. They will listen on 437.125, 437.175, 437.225, 437.275, 437.325, and 437.375 MHz; transmit on 145.450, 145.475, 145,550, or 145.575 MHz. 145.575 down, 437.275 MHz up will be the normal pair.

Rumors Of 10 Or 15-Meter Operation Still Persist at presstime; an operating schedule for DPØSL has not yet been released.

comments

practically speaking

Dear HR:

Joe Carr, K4IPV, is off to a fair start with his new column, "Practically Speaking." The first (September, page 67) was good reading. I have to disagree with him, however, when he says, "The first time to think about repairs is while you're unpacking the new rig."

Wrong.

The first time to think about repairs is *before* the rig is bought. The unpacking stage is too late.

Before plunking down several kilobucks for one of the so-called state-of-the-art transceivers being marketed these days, ham operators need to give serious, careful thought to the crucial question of servicing and maintaining the rig(s) under consideration. Is that aspect of ownership going to cost an arm, a leg and another part of the anatomy — to say nothing of the time spent without the rig while it is being repaired?

Are all those gongs, whistles, and bells *really* worth it? Does one truly need rigs that scan, memorize, and do all sorts of other gimmicks?

Two years ago two friends here returned from a national Amateur Radio convention, each with a new rig. One, who purchased the latest offering from Brand X had to return the rig to the factory once while it was still under warranty and twice since then. He is no slouch at troubleshooting or repair, but the complicated circuitry was more than he could handle. Labor, parts, and shipping charges thus far have totaled more than one-third the original cost of the transceiver, and he has been without it almost 60 days altogether.

My second friend brought home a shiny, new Brand Z transceiver. It was almost a pure vanilla rig with no gong, whistles, or bells. It has sat in his shack and has worked, day in and day out. What few repairs have been needed were done quickly, easily and inexpensively. He said he decided on that rig because of the ease and simplicity of operation.

Practically speaking, which do you think was the more practical?

Fred Conavita, W5QJM
Austin, Texas

audio filter design

Dear HR:

Instead of transforming the normalized values of a five-branch Butterworth low-pass prototype design to get a five-branch high-pass as discussed in Stefan Niewiadomski's article, "Passive Audio Filter Design," part 2, October, 1985, a simpler procedure would be to use standard-value capacitor (SVC) filter tables to select an appropriate design. The fact that the SVC capacitor design values are exactly identical to the commercially available standard values simplifies construction. For example, a "near-Butterworth" 500-ohm design with a VSWR of 1.023 and a 3-dB cutoff frequency of 487 Hz (within 3 percent of the desired 500 Hz cutoff frequency) is available with standard values of C1,5 = 0.82 μF and C3 = 0.33 μF. L2,4 = *96.2 mH.* The 20-dB and 40-dB attenuation frequencies are 330 and 214 Hz. Some SVC filter tables appear in *The ARRL 1985 Handbook, (pages 2-40 through 2-44)* and a more complete selection will appear in the next edition of the *Radio Handbook*, edited by Bill Orr, and to be published soon by Howard W. Sams and Co., Inc. The SVC filter tables were also published in the record of the IEEE 1985 International Symposium on Electromagnetic Compatibility held in Wakefield, Massachusetts from August 20-22. *ham radio* readers should therefore understand that the preferred procedure for designing simple passive LC filters, such as generally required by the Radio Amateur, is to use SVC filter tables. The procedure explained by Niewiadomski should be used only when there is a special requirement for a specific response type and a precise cutoff frequency.

Ed Wetherhold, W3NQN
Annapolis, Maryland

J-pole or Zepp?

Dear HR:

K1WWT is right, (see "J-pole or Zepp," Comments, *ham radio*, February, 1985, page 8). The J-pole antenna described in KD8JB's earlier *ham radio* article* is a Zepp antenna.

The Zepp antenna, named for the Zeppelin airship, on which it was originally used, is not a 3/4-wave *antenna* operating against a 1/4-wave counterpoise as KD8JB mentioned in the reply comments.

The classical Zepp antenna, now more than 75 years old, is a full-wave current-fed antenna with a 1/4-wave section folded over on itself. This leads to a 1/2-wave end-fed radiator with 1/4-wave matching transformer. The antenna at first was a balloon antenna (**fig. 1**) and looked like an inverted J-pole dangling from the airship. In the beginning the matching transformer and the radiator hung in a straight line. Later on the matching transformer was set at right angles to the radiator. This low-voltage input feed arrangement was a remarkable improvement over the dangerous practice of using a high-voltage feed in the presence of the oxyhydrogen-gas with which the balloons were filled.

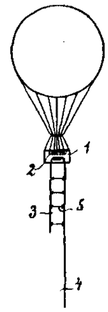
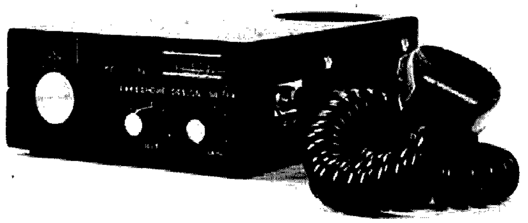


fig. 1. The Zepp antenna, designed by Dr. Hans Beggerow, was patented in Germany in 1909. The illustration above accompanied the original patent.

Alois Krischke, DJ0TR/OE8AK
Munich, West Germany

*See "All-metal, 2-meter J-pole Antenna," *ham radio*, July, 1984, page 42.



a compact
75-meter monoband transceiver

Modular design yields 30 watts PEP *and* high performance

This article describes a compact monoband SSB transceiver that employs broadband techniques, IC building blocks, and an FET power chain. A detailed block diagram that shows all module interconnections is shown in **fig. 1**. As an extension of an earlier receiver project, the design provides all of the basic features required for convenient operation.[1] The receiver section offers excellent sensitivity and selectivity, audio-derived AGC, an S-meter, headphone or speaker operation, and above-average audio quality. The transmitter has amplified ALC and delivers 30 watts PEP to a 50-ohm load. The completed package is about the size of a 2-meter FM transceiver, measuring 2 × 5 × 6 inches (5 × 12.7 × 15.25 cm) and weighing about 2 pounds (1 kg).

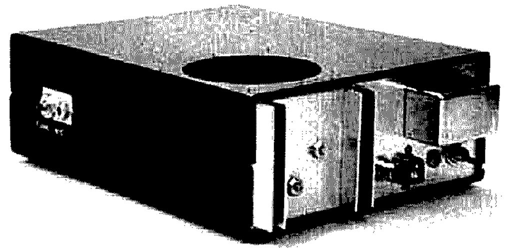
circuit description

The transceiver employs a single conversion frequency plan with a 9-MHz IF and a 5.0-MHz VFO. Receiver preselection is provided by a two-section bandpass filter (see **fig. 2**). Additional HF rejection is obtained from the transmitter's low-pass filter. Receiver mixer U1 is an active DBM which has been biased for maximum gain. Mixer output is fed to crystal sideband filter FL1 through a simple diode switching network.

IF stage U2, shared by the transmitter, provides 45 dB of gain with an AGC range of about 70 dB. Gain for the entire receiver is controlled via U2's AGC line. Automatic control is audio-derived from the output of audio amplifier U4. Manual control is provided by a voltage divider circuit. During the receive cycle, the two control voltages are gated onto the AGC line through diodes. The output of IF amplifier U2 is simultaneously fed to product detector U3 and transmit mixer U7. U3, an active DBM product detector, provides audio detection and additional system gain.

Since gain is controlled exclusively through IF amplifier U2, audio amplifier U4 operates at full gain (see **fig. 3**). U4 provides 400 mW of output — more than enough to drive the transceiver's small built-in speaker. Attenuation is provided for speaker protection and for headphone operation.

AGC voltage is sampled from the output of U4, detected, and fed to DC amplifier Q1. The RC time constant of Q1 is switched to provide slow release time during receive, and fast release time during transmit. Q2 provides additional amplification of the control signal, sets the AGC threshold for U2, and drives meter M1. M1 functions as an S-meter during receive and as an ALC indicator during transmit. The entire receiver section operates from a 12-volt source with an average current drain of only 50 mA on receive.



A small heatsink is sufficient for intermittent SSB operation, but area should be increased for CW operation. Mounting FL1 on rear panel saves internal space.

By Rick Littlefield, K1BQT, Box 114 Barrington, New Hampshire 03825

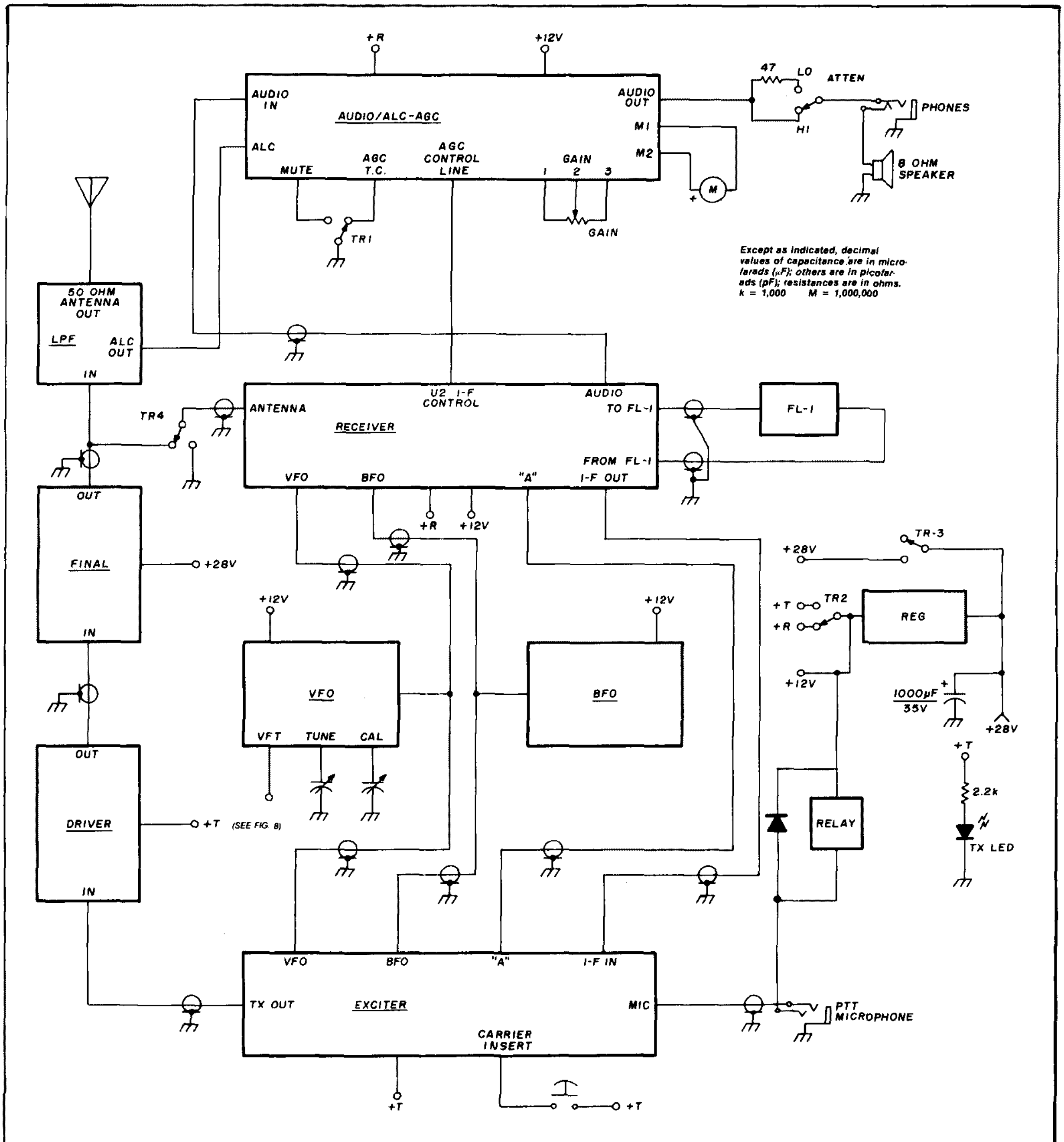


fig. 1. Module interconnections

Low-Z microphone amplifier U5 is a standard op-amp circuit which develops the necessary audio voltage to drive balanced modulator U6 (see fig. 4). Like all other mixing devices in the transceiver, U6 is an active DBM. Provisions are made to unbalance the device when carrier is needed for RF chain or antenna tuner adjustment. The output of U6 is fed through a diode switching network to sideband filter FL1 and IF amplifier U2. As noted earlier, ALC voltage is applied to U2 during transmit to maintain high transmitter output without driving the RF chain into saturation.

Transmit mixer U7 combines the IF signal from U2 with VFO drive to produce 75 meter output (a CW-only design would substitute BFO drive for the IF signal). The output of U7 is buffered and amplified by

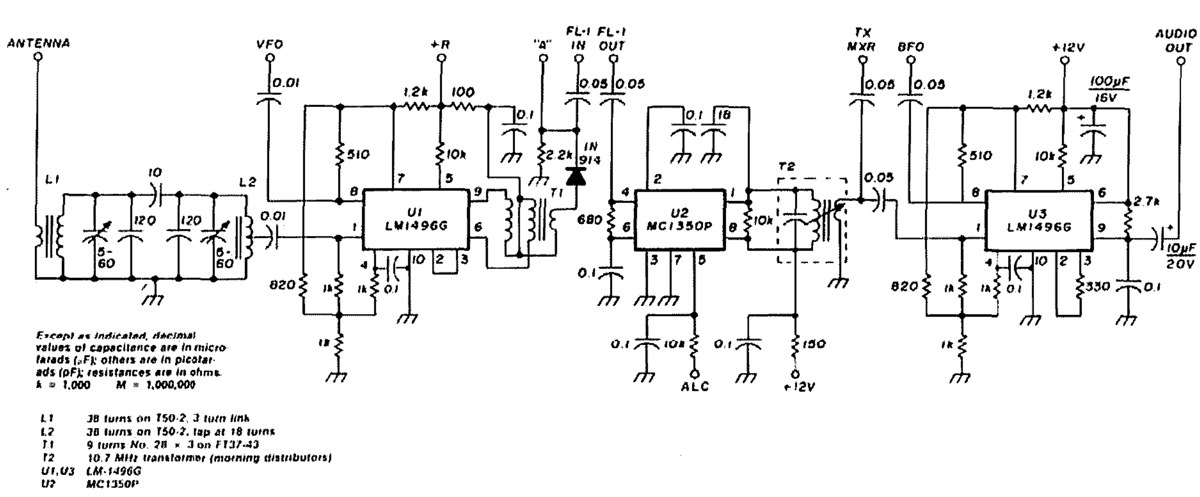
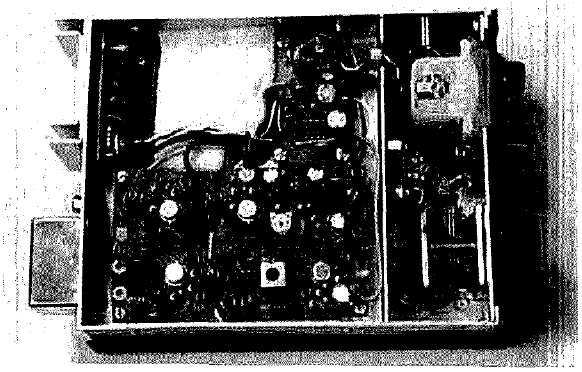


fig. 2. Receiver module. All phasing and combining of transformer leads is done via the circuit board layout. Transformer leads are simply inserted as shown.



Receiver, exciter, and audio modules are located in the top compartment of the cabinet. VFO is in a separate front compartment.

pre-driver Q3 (see **fig. 5**). If CW operation is anticipated, keying can be added to this stage by simple modification. Output can be reduced for QRP operation by adjusting the bias on gate No. 2. The output of Q3 is filtered by a two-section bandpass filter. Driver Q4, an inexpensive HEXFET, develops 300 mW PEP — enough power to drive PA Q5. A three-element low-pass filter reduces harmonic content prior to final amplification

The final amplifier Q5 is a Motorola T-MOSFET operating in class AB (see **fig. 6**). At 4 MHz, this 28-volt device operates at approximately 70 percent efficiency, provides 20 dB of gain, and delivers 30 watts PEP into a 50-ohm load. The output of Q5 is transformed to 50 ohms through 4:1 balun T5 and fed into a 7-element Chebyshev low-pass filter. A diode detec-

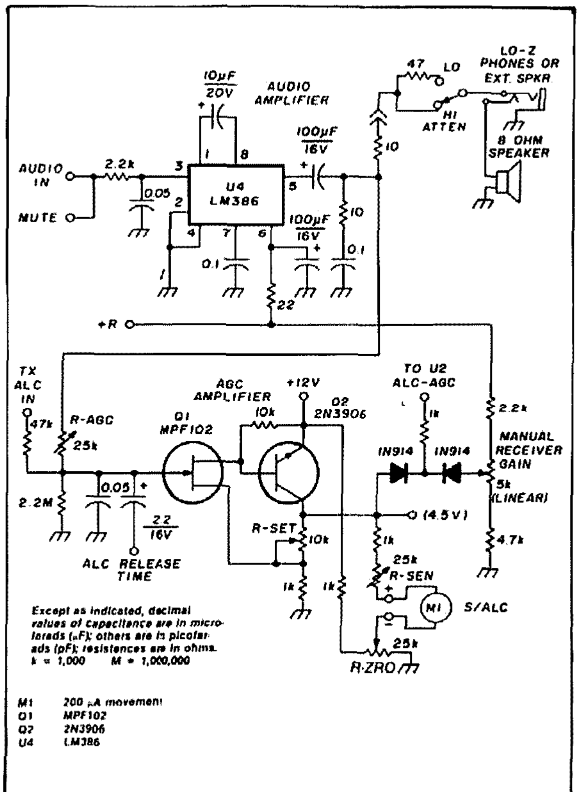


fig. 3. Audio amplifier/ALC-AGC.

tor at the LPF input samples amplifier output for ALC (see **fig. 7**).

All transmit and receive mixing functions are handled by active DBMs, devices which require no more

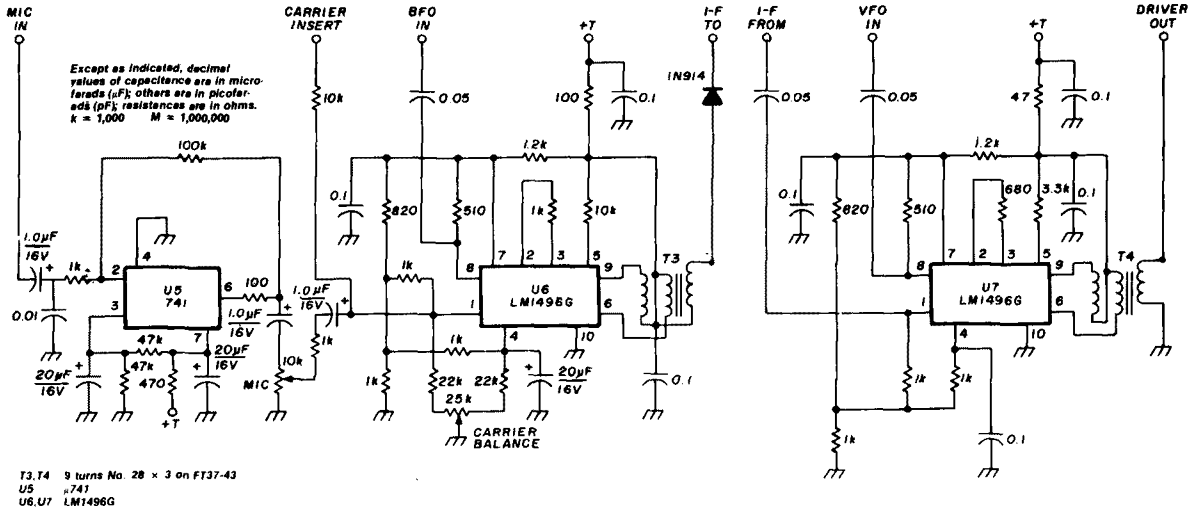


fig. 4. Exciter module.

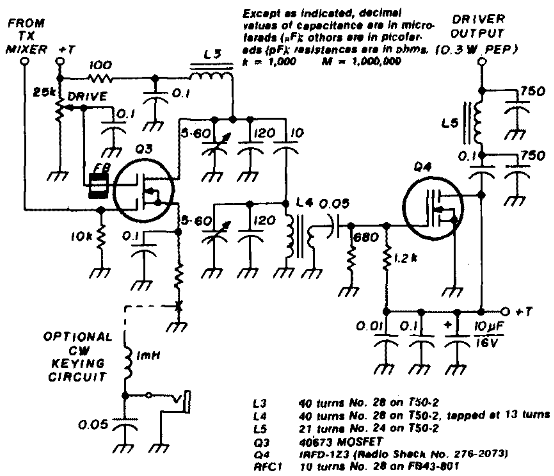


fig. 5. Driver/BPF module.

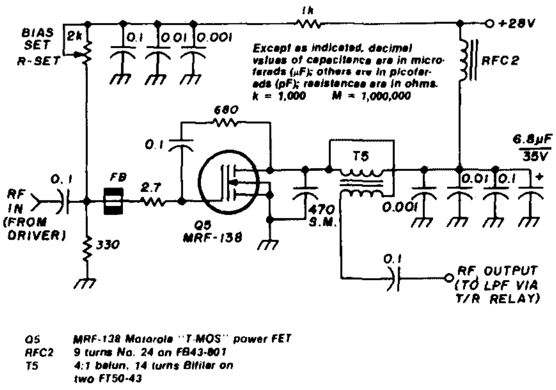


fig. 6. Final amplifier module.

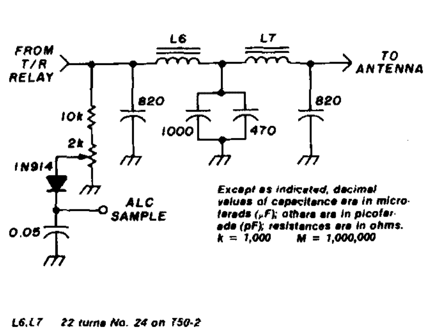


fig. 7. LPF/ALC detector module.

than 100 mV of injection. Consequently, drive demands on the VFO and BFO are minimal. VFO Q6 is a popular Hartley JFET circuit which is buffered by source follower Q7 (see fig. 8). An optional VFT control aids fine tuning. BFO Q8 is a simple unbuffered crystal oscillator (see fig. 9). If the transceiver is modified for 20 or 15-meter operation, the BFO output should be carefully matched to its load and filtered for harmonic content.

The PA is the only stage requiring a 28-volt source. An on-board monolithic voltage regulator reduces supply voltage to the other stages. The compact 28-volt

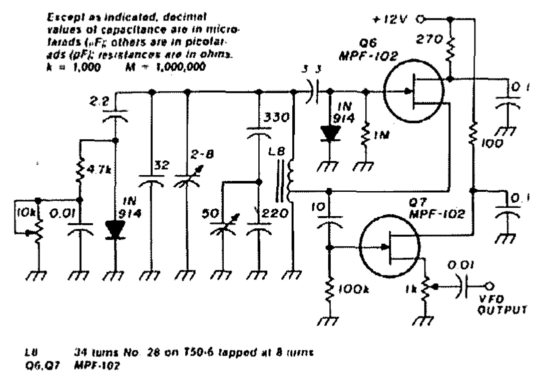


fig. 8. VFO module.

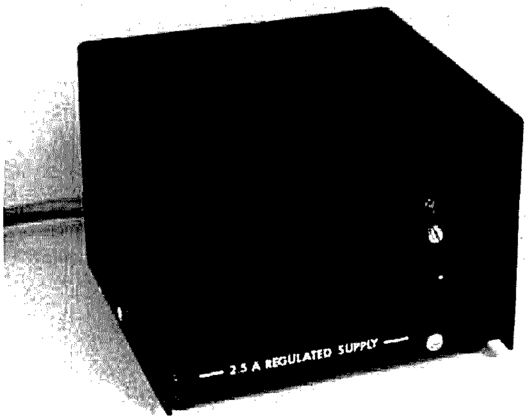


fig. 10. The transceiver's compact power supply measures only 4 × 3 × 3 inches (10 × 7.6 × 7.6 cm). The design is simple, and can be built entirely from off-the-shelf Radio Shack components.

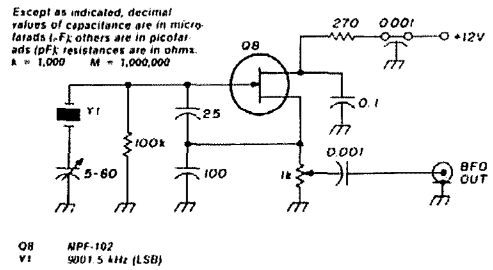


fig. 9. BFO module.

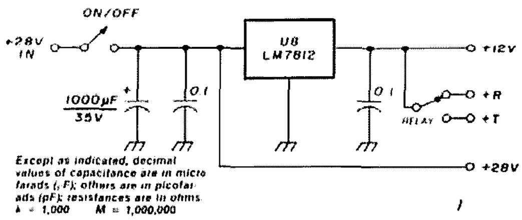


fig. 12. Transceiver on-board regulator.

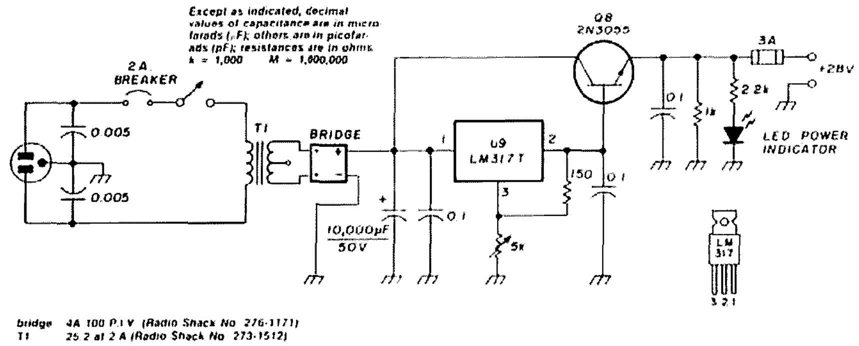


fig. 11. External power supply schematic.

external power supply and schematic shown in **figs. 10** and **11**, respectively, was built from off-the-shelf Radio Shack components. The output of transformer T1 is bridge rectified, filtered, and regulated by pass transistor Q1. Adjustable regulator U1 drives the base of Q1 to set output voltage and to provide additional electronic filtering. An on-board regulator is also incorporated in the transceiver (**fig. 12**).

construction

The boards for this project were laid out in modular strips to facilitate modification during the design pro-

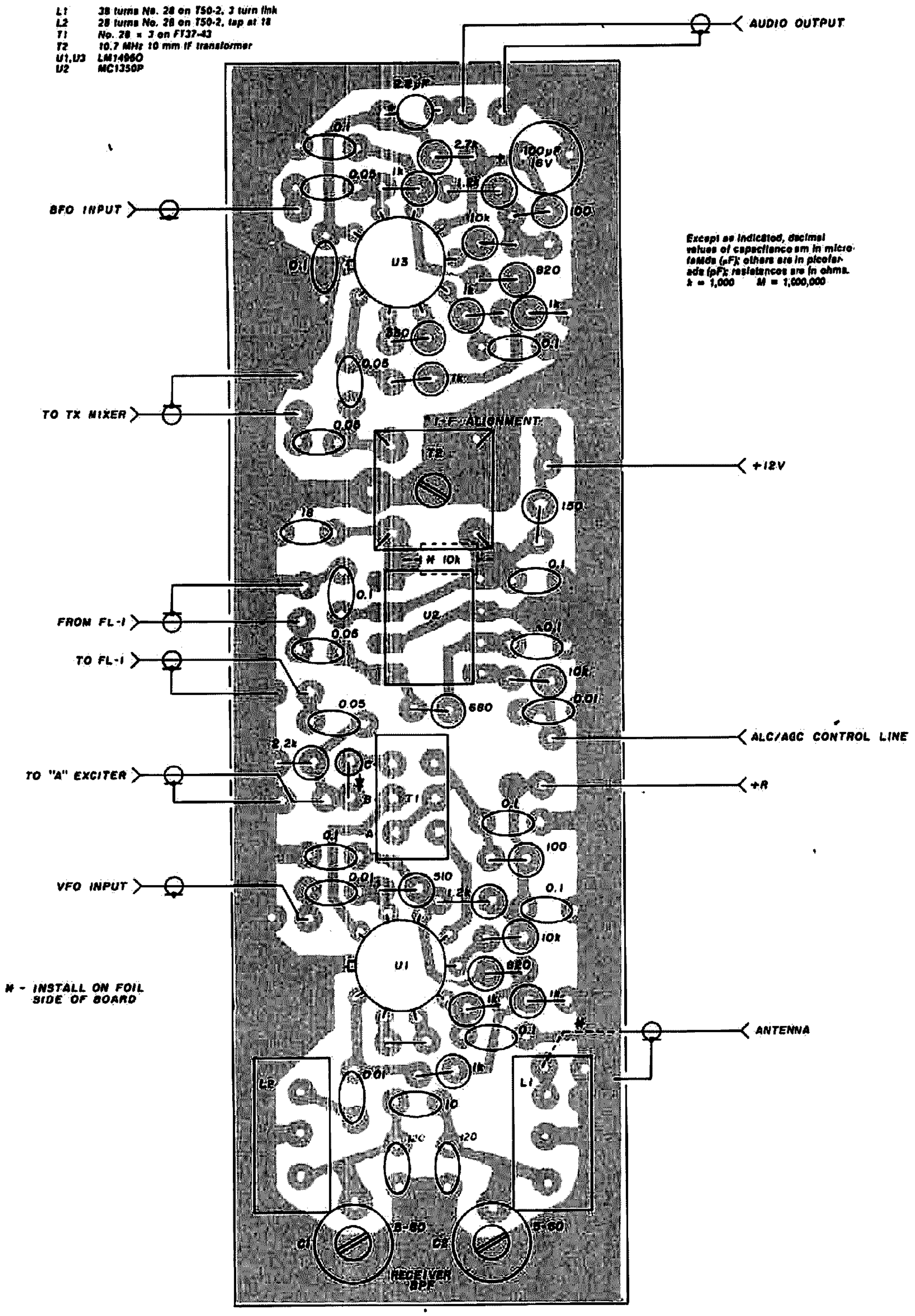


fig. 13. Receiver module.

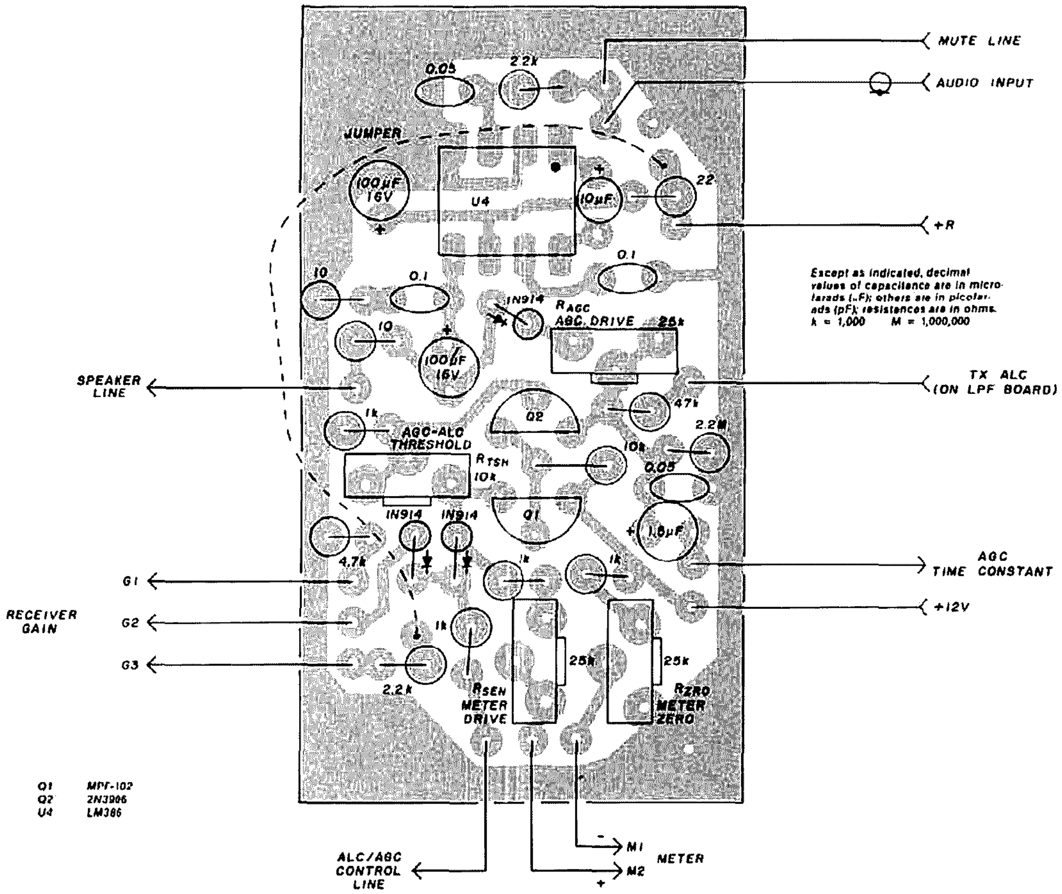
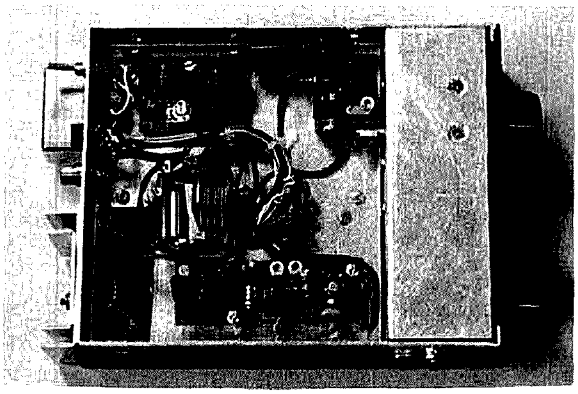


fig. 14. Audio/AGC module.



Bottom view shows BFO, driver, and low-pass filter modules. Voltage regulator is mounted on the side of the case. A 4PDT relay for T-R switching simplifies circuit design and reduces receiver power consumption.

cess. To make the job of interstage wiring easier, the exciter and receiver strips were later joined together.*

Figures 13 (receiver module), **14** (audio/AGC module), **15** (exciter module), **16** (driver/bandpass filter), **17** (MRF-138 final amplifier module), **18** (low-pass filter board), **19** (VFO module), and **fig. 20** (BFO module) show the printed circuit board patterns and component layouts.

Board assembly is routine, but a few specific points deserve mention. The boards were designed around miniature parts. Substituting 1/2-watt resistors, high-voltage capacitors, and other large components can quickly result in overcrowding. Since the parts density is quite high, double-checking parts placement against the schematic or a layout is also recommended.

Use care when winding toroidal transformers and chokes. Most FT (ferrite) cores have rough edges that can easily abrade the insulation from enamel covered

*A complete kit containing all parts, etched pre-drilled circuit boards, punched, painted enclosure, and assembly manual is available from Radiokit, Box 411, Greenville, NH 03048

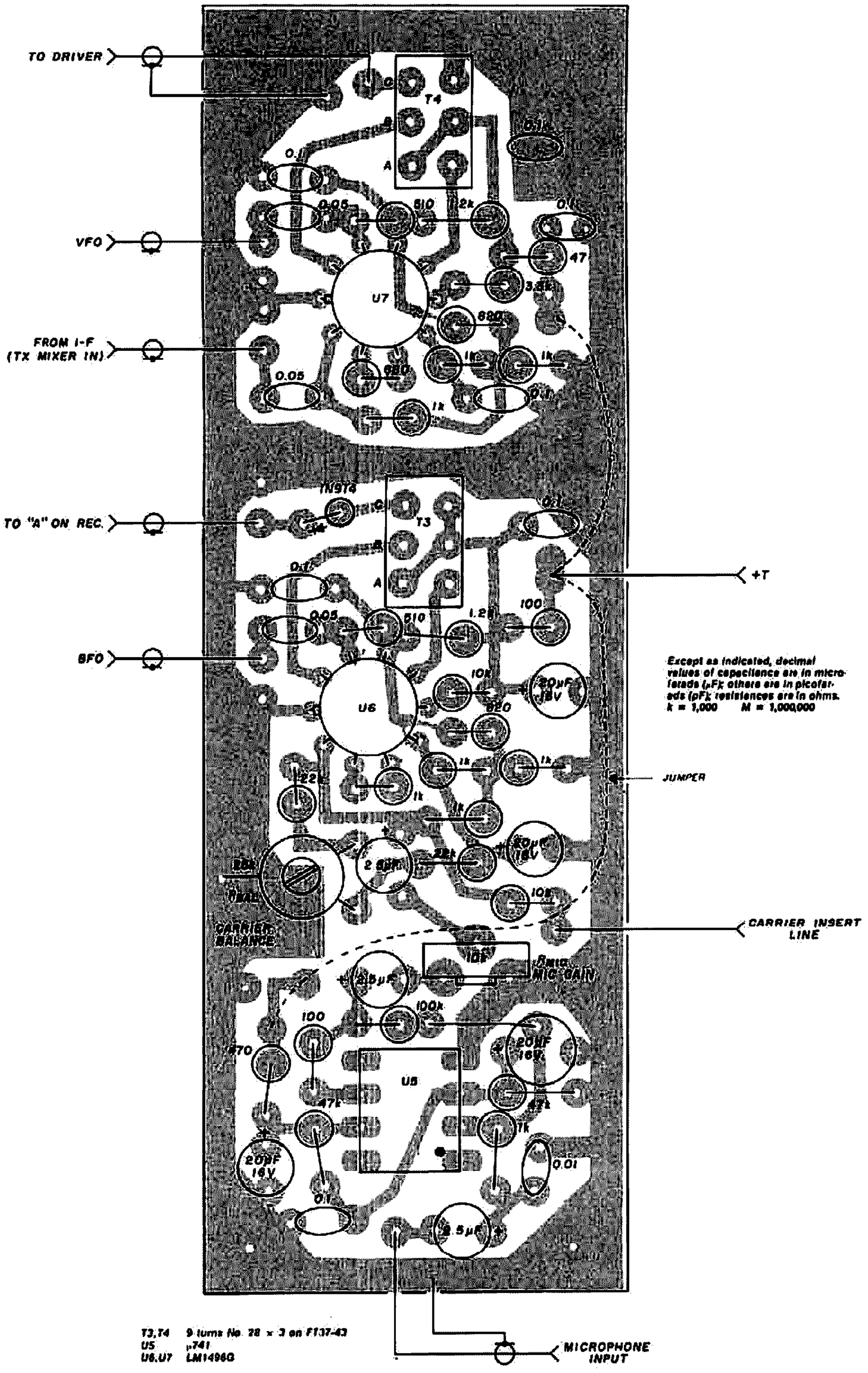
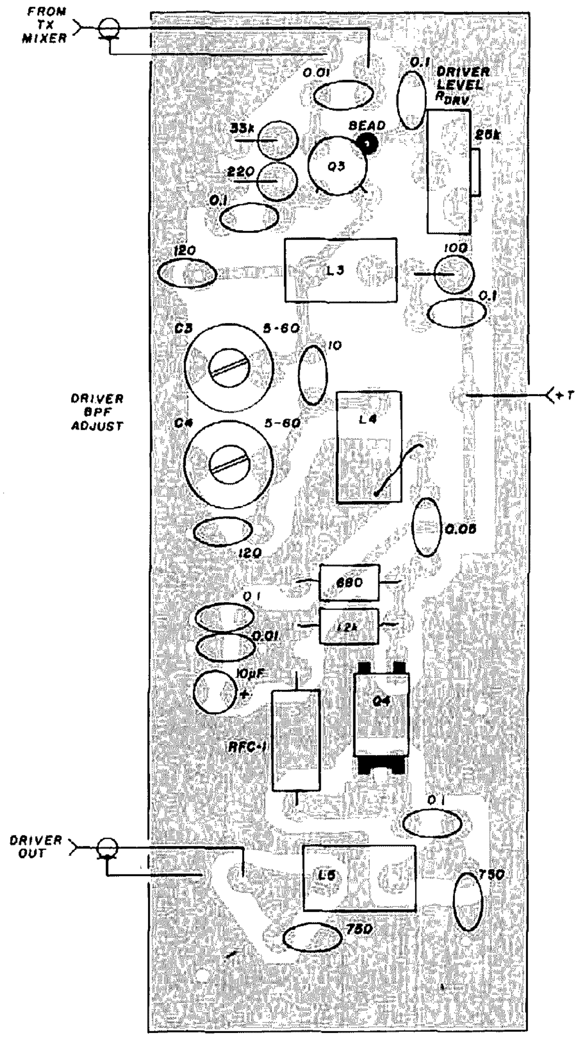


fig. 15. Exciter module.



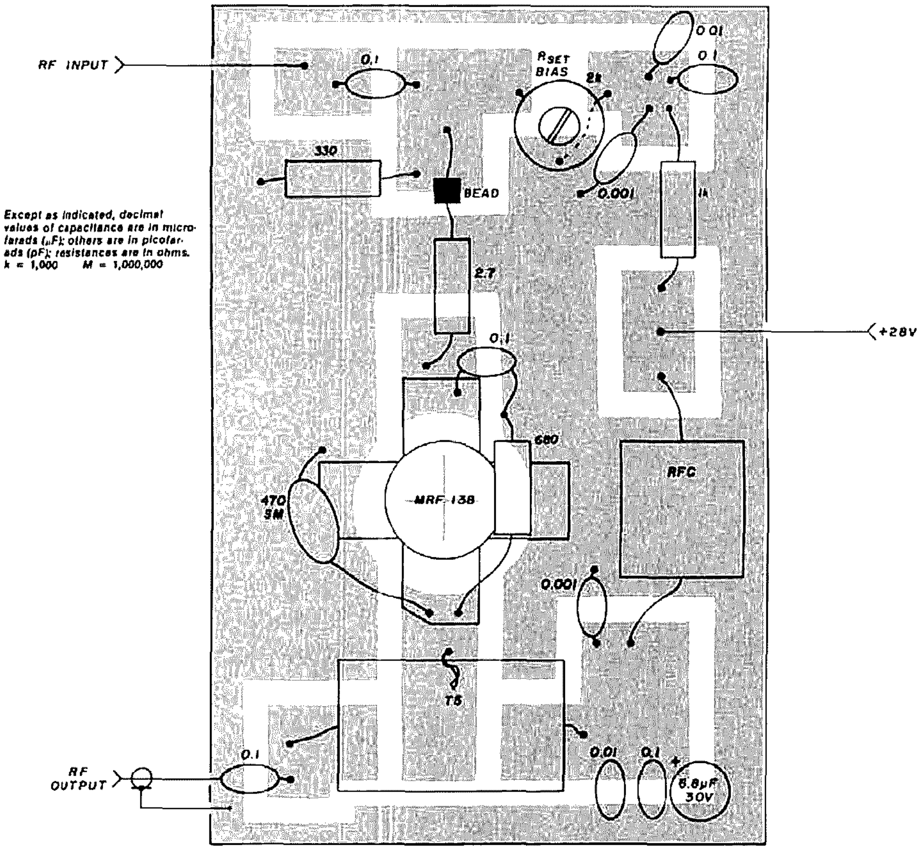
Except as indicated, decimal values of capacitance are in micro farads (µF); others are in picofarads (pF); resistances are in ohms.
k = 1,000 M = 1,000,000

L3	40 turns No. 28 on T50-2
L4	40 turns No. 28 on T50-2, tap at 13
L5	21 turns No. 24 on T50-2
Q3	40673
Q4	IRFD-123
RFC-1	10 turns No. 28 on FB43-801

fig. 16. Driver/bandpass filter.

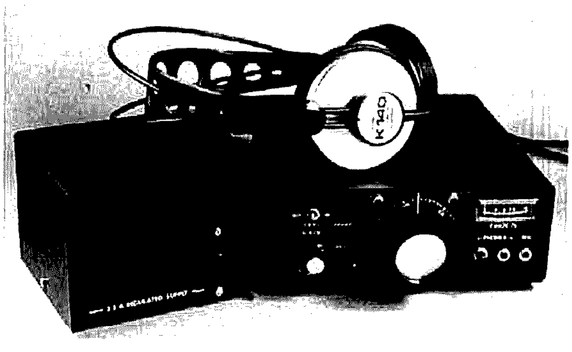
wire. Prepare all FT cores in advance by smoothing the corners and applying two coats of clear nail polish ("T cores are usually epoxy coated and require no preparation). Since most toroid devices have delicate leads that are easily broken, they should be mounted last and glued securely to the board with Ambroid™ cement.†

†T2, a 10.7 MHz 10 mm green-core transformer, is available from Morning Distributing Co., P.O. Box 717, Hialeah, Florida 33011.



RFC 8 turns No. 24 on FB43-801
T5 14 turns No. 24 Bifilar on two FT50-43

fig. 17. MRF-138 final amplifier.



The transceiver can be packaged to suit the builder. The larger cabinet shown here provides more panel space and room for a larger speaker.

Because the VFO circuit is vulnerable to thermal and mechanical instability, its construction requires special attention. These hints should be followed to insure satisfactory performance:

• Use NPO capacitors in the tank circuit wherever possible — the thermal characteristics of silver mica capacitors are often unpredictable.

• To immobilize VFO inductor windings, dip or paint the coil in clear nail polish and sandwich it to the board between fiber washers with a non-inductive screw.

• To insure smooth tuning, the main VFO variable should be a ball-bearing design with either a built-in or external reduction drive.

• The calibration trimmer should be a miniature air-variable (ceramic trimmers drift).

• When installing the VFO module, make all leads connecting external VFO components as short and rigid as possible.

• To avoid thermal instability, locate the VFO away from heat-generating stages such as the PA and the voltage regulator.

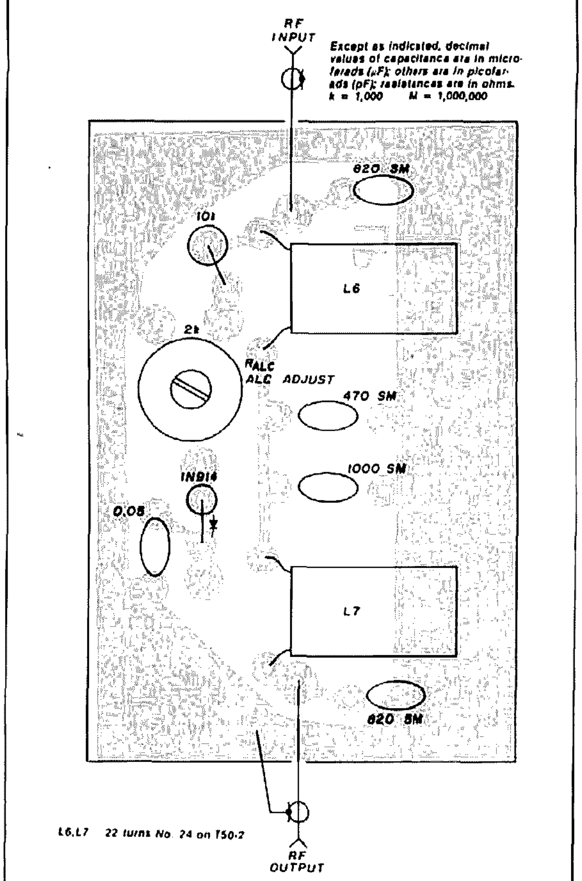


fig. 18. Low-pass filter board.

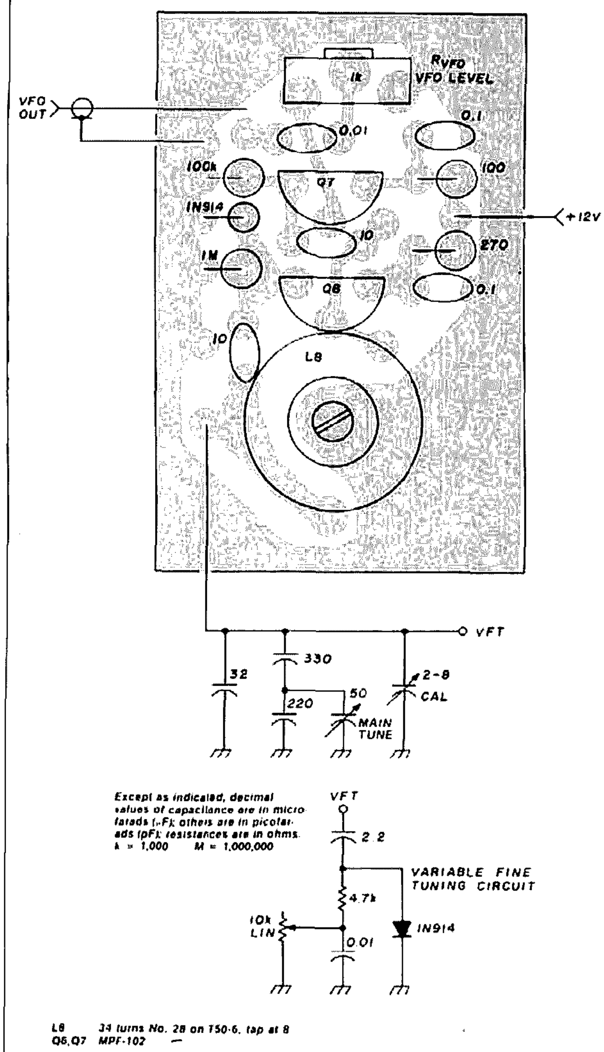


fig. 19. VFO module.

- Shield the VFO compartment against strong RF fields.
- Locate the transceiver's T-R relay away from the VFO compartment. If located too close, the relay's magnetic field will produce an unwelcome frequency shift.

The only other stage requiring special care during construction is the final amplifier. To prevent the possibility of VHF parasitics, strip-line construction is used and components are soldered directly to the top of the board. To insure strong solder connections, each lead should have a short 90-degree bend at its end in order to make flat contact with the board's surface. The transistor should be mounted first. The MRF-138 is an unprotected MOS device, and I recommend using a grounded iron and wrist-strap to prevent static build-up during installation. Once the module is completed, the circuitry will protect the device.

The transceiver cabinet is a bi-level design fabricated in a custom sheet-metal shop. While this packaging contributes to the appearance and small size of the finished unit, much simpler cabinetry is perfectly acceptable as long as a few basic conditions are met. First, all recommendations to insure VFO stability should be observed. Second, the PA module should be mounted to the inside of the back panel with a suitable external heatsink provided on the opposite side. Additional heatsink area is required for extended phone, RTTY, or CW operation. Finally, the VFO and BFO should be fully shielded.

In any modular project, interstage wiring can become a nightmare if the wire is prone to breakage or is difficult to handle. Selecting only highly flexible small-diameter wire and shielded cable keeps interstage harnesses small and manageable. I have found that lavalier microphone cable is smaller and much

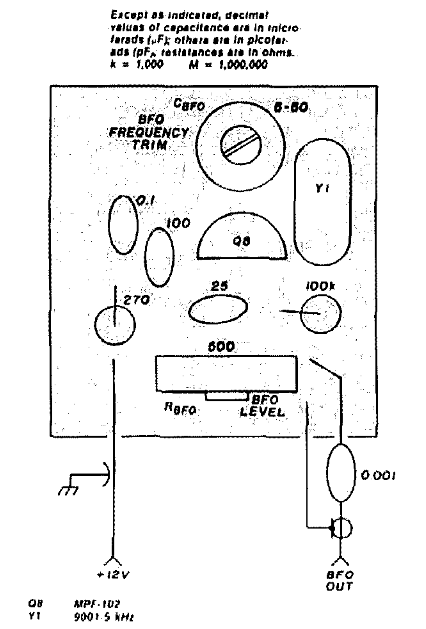


fig. 20. BFO module.

easier to handle than RG-174 miniature coax. Mounting boards on 1/4-inch standoffs also contributes to a neat layout, since this provides space for interstage wiring to pass underneath. Once the modules are mounted and interstage wiring is completed, testing and alignment can begin. Power distribution and T-R switching should be thoroughly checked first, since an error here could damage components.

alignment

Fixed capacitor values in the VFO tank may require some substitution to establish the desired operating range (5000 kHz to 5200 kHz for 4000 to 3800 kHz operation). Once this range is established, a tuning dial can be calibrated. A frequency counter facilitates the calibration process. Once the VFO dial is calibrated, receiver alignment can proceed. Use **fig. 21** to locate the calibration and alignment controls.

- Connect the receive and transmit mixers to the VFO, and adjust R_{VFO} for 100 mV RMS output.
- Connect the product detector and balanced modulator to the BFO, and adjust C_{BFO} for an operating frequency of 9001.5 kHz. Adjust R_{BFO} for an output of 100 mV RMS.
- Set the receiver AGC threshold by adjusting R_{TSH} for 5 volts as measured at TP1. Zero the S-meter via R_{ZRO}.
- Set the receiver gain fully clockwise for maximum gain and adjust IF transformer T2 for a peak in background noise.
- Connect a 50-ohm antenna and tune the VFO to 3900 kHz. Peak bandpass filter trimmers C1 and C2 for maximum sensitivity.

The receiver should now be fully functional. Check AGC action by tuning in an extremely strong SSB signal. If the audio cracks and distorts at full gain, the AGC is under-controlling IF stage U2. Increase AGC gain via R_{AGC} to eliminate this condition. If the audio "pumps" on voice peaks or motorboats with no signal, the opposite conditions exist and AGC gain should be decreased. Meter sensitivity control R_{SEN} should be adjusted so that extremely strong signals register in the upper 10 percent of the scale.

To prepare for transmitter alignment, disconnect the 28-volt supply line from the final amplifier board. Terminate the output of the driver with a 47-ohm resistor and connect a scope across the termination. Microphone gain R_{MIC} should be set fully off, and predriver gain R_{DRV} set to the middle of its range (maximum gain). Tune the VFO to 3900 kHz.

- Key the transmitter and activate the carrier insert switch. Adjust IF transformer T2 and bandpass filter trimmers C3 and C4 for maximum output.
- Key the transmitter and adjust R_{BAL} for minimum carrier output. A receiver tuned to the output frequency may provide a better null indication.
- Connect a 500-ohm dynamic microphone and advance R_{MIC} to 75 percent. Speak into the microphone and watch the scope for signs of instability ("grass" or parasitic oscillations on the waveform). The pattern may show flat-topping on voice peaks, since the ALC is not yet functional.

If instability or parasitics are observed, find their source before going on. Check the RF amplifiers in isolation, and check IF amplifier U2 (reducing the value of the 10-kilohm resistor across the primary of T1 should tame unstable operation in U2). If operation is normal through the driver stage, alignment can continue.

- Connect the 28-volt supply line to the final amplifier board through an ammeter. Short the amplifier's input terminal to ground. Key the transmitter, and adjust R_{SET} for an idling current of 250 mA. Note that this adjustment is sensitive to changes in supply voltage. If the power supply voltage is changed significantly at a later date, the bias should be re-set.
- Remove the driver termination, unshort the input to the PA, and hook up the driver. Connect a 50-ohm dummy load to the output of the transceiver. Place a single turn pick-up loop through balun T5, and connect it to the scope. The driver low-pass filter and bandpass responses are shown in **figs. 22A, B**.

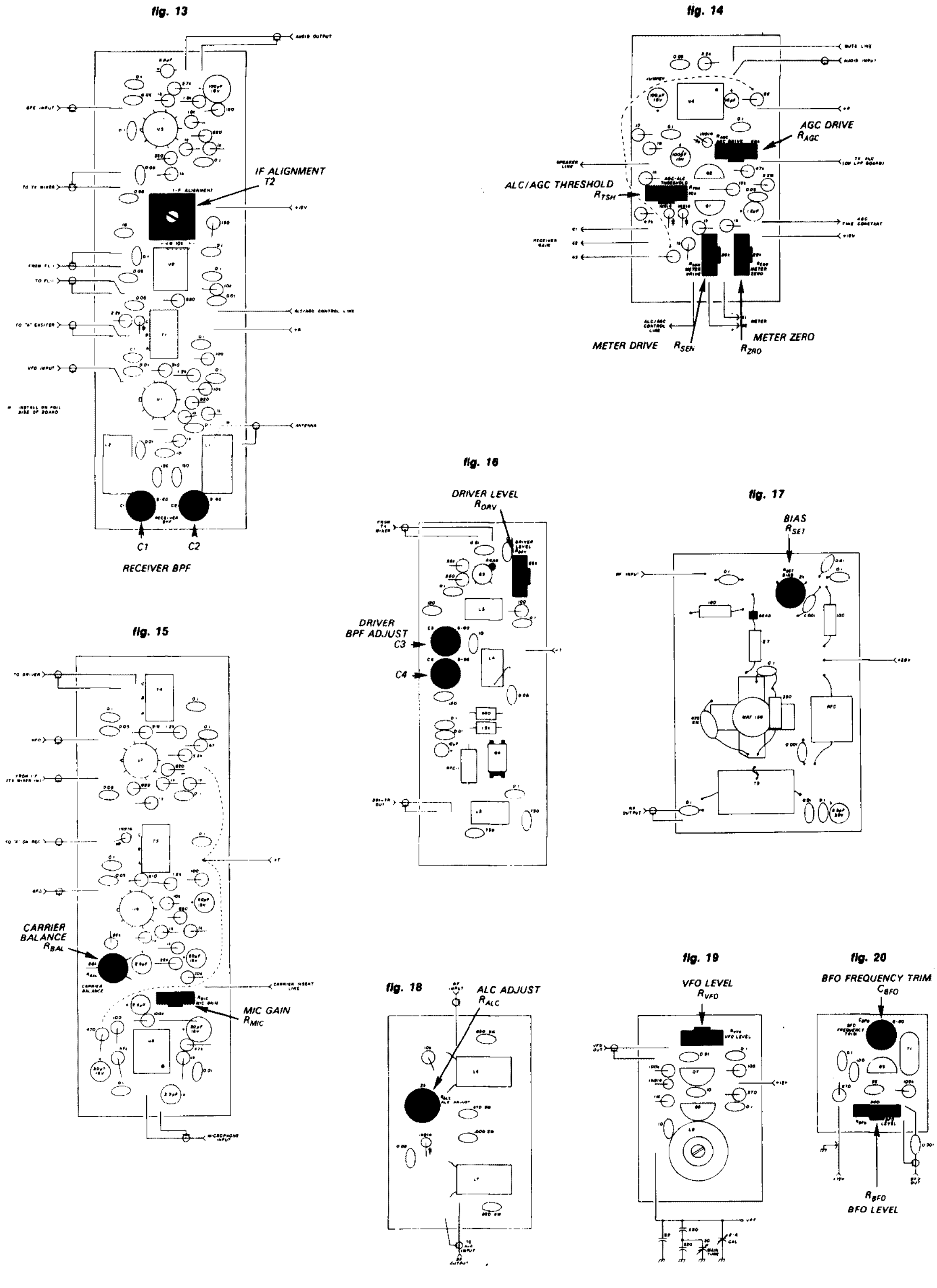


fig. 21. Calibration and alignment controls.

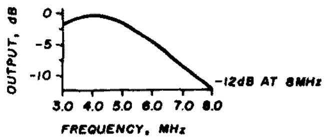


fig. 22A. Driver low-pass filter response.

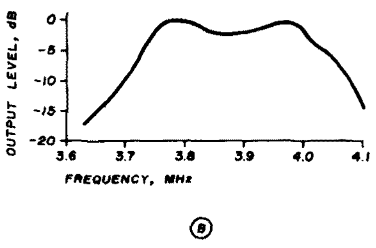


fig. 22B. Driver bandpass filter response.

- Key the transmitter and speak loudly into the microphone, adjusting R_{ALC} for maximum transmitter output. The scope pattern should show flat-topping on voice peaks. If saturation does not occur, increase microphone gain until it does. If the final cannot be saturated, system gain is low and a problem exists (check the AGC threshold voltage at TP1 first; if set above 5 volts, transmitter gain is reduced).

- To set ALC level, close-talk the microphone and adjust R_{ALC} to the point where flat-topping just disappears. The ALC meter should deflect past mid-scale on voice peaks and a power meter should indicate an average output of 10-15 watts.

This completes transceiver alignment.

performance

The transceiver was tested to see if performance approximated industry standards and met FCC regulations for spectral purity. Receiver noise floor was measured at −120 dBm. Selectivity reflected the published specifications of filter FL1. AGC held a 60 dB change in signal strength to a 3 dB change in audio output. AGC attack was a bit slow, resulting in some audible "cracking" on extremely strong signals. This condition is not uncommon in simple audio-derived systems. Overall receiver audio quality was judged excellent when compared against a popular imported multi-band transceiver. Tests for receiver intermodulation distortion were not conducted.

At 30 watts PEP output, transmitter IMD was measured at −30 dB. Second and third harmonics were −47 dB and −55 dB, respectively. Saturation occurred at 35 watts PEP. Transmit audio reports were generally excellent, but microphone selection was an important factor. Low-Z broadcast dynamics produced the best overall quality, but an inexpensive mobile microphone provided a bit more "punch" under difficult band conditions. The MRF-138 final amplifier survived open and shorted port conditions without damage, indicating acceptable immunity to high SWR.

operation

The transceiver's small size makes it a natural for traveling, or for use as a second station at home. Mine resides in a corner of the family room on a small writing desk, close to the wood stove, kitchen, and other comforts of home. On-air performance has been very gratifying. Using an inverted-V antenna at 50 feet, I have worked all U.S. call areas, operated contests, controlled nets, and elbowed my way through evening QRM with excellent regularity. In evaluating the transceiver's effectiveness, it is important to remember that dropping transmitter output from 100 watts to 30 watts reduces the received signal less than 1 S-unit. Under most band conditions, this is not significant.

conclusion

My goal was to design and build a simple monoband SSB transceiver that would be compact, easy to replicate, and powerful enough to provide reliable communication on 75 meters. Off-the-shelf components and contemporary design techniques were employed wherever possible to make the job easier. The transceiver described in this article is my third, and carries with it the experience of the first two. With minor modifications, the design should be transferable to other bands. I hope this article will encourage others to take the plunge and build — there's no magic involved, and the enjoyment that comes from operating a homebrew rig is fantastic.

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ham radio

high stability local oscillators
for microwave receivers and other applications

Phase-locked approach offers greater stability

This article describes a reliable phase-locked loop originally designed for the first conversion of a radio astronomy microwave receiver[1] applicable for general microwave receiver use and to other HF, VHF, and UHF phase-locked applications as well.[2] Its design evolved from one of my earlier efforts incorporated in a fully synthesized, general coverage HF transceiver as shown in **fig. 1**.

Local oscillators used for conventional microwave (TVRO) such receivers are usually open loop and are installed outdoors as part of the first converter, known as the "head end" and located at the feed point of a parabolic dish antenna. They normally exhibit gross frequency instability (typically ± 1.5 MHz) due to their free running characteristics, which are affected by ambient temperature changes as well.

The two types of open-loop local oscillators most commonly used for this application are the free running tuned cavity and the crystal controlled multiplier type. This article deals with the second approach, which allows an already-clean multiplier chain to lock on a much more stable reference frequency strategically located away from the elements. Consequently, the unit can be used under varying temperature conditions and will follow a remote reference source kept indoors for good stability.

The synthesizer uses a simple version of a type-one crystal-controlled phase-locked-loop operating at 88 MHz.[3] Its output is divided by a fixed modulus divider involving emitter-coupled-logic (ECL) and transistor-transistor-logic (TTL) circuits. The division number N is fixed at 88 and followed by an exclusive OR digital phase detector that closes the loop through a simple low-pass filter as shown in **fig. 2**.

In order to obtain the desired microwave frequency, the output of the oscillator running at exactly 88 MHz is used to drive a times-12 multiplier such as described by Paul Shuch, N6TX.[4] Other multiplications are possible for even higher frequencies. The reference frequency can be supplied to the synthesizer from the back end of the receiver via fiber optics or coaxial cable communication links (depending on the distance; digital line drivers may be required in the latter). With this approach, a remotely located temperature-compensated crystal oscillator (TCXO) that acts as a time base will maintain the short and long term frequency stability of the 88-MHz crystal oscillator through the phase-locked technique. The stability of the multiplier chain in the microwave receiver will thereby also be favorably affected. High initial stability and spectral purity would be required to compensate for the magnifying effect of the multiplier. My circuit used a 4-MHz TCXO manufactured by McCoy Electronics for the reference oscillator. This part guarantees $\pm 5 \times 10^{-7}$ (± 0.5 PPM or ± 2 Hz at 4 MHz) Hertz per year.[2] This represents an ultimate stability for the remotely located L-band local oscillator of ± 6 Hertz, respectively.

The circuit design of the synthesizer is simple (although making one work is another story) as shown in **fig. 3**. A highly stable 88-MHz (0.001 percent) fifth overtone crystal was chosen to guarantee initial start-up almost on frequency before locking occurs. It is

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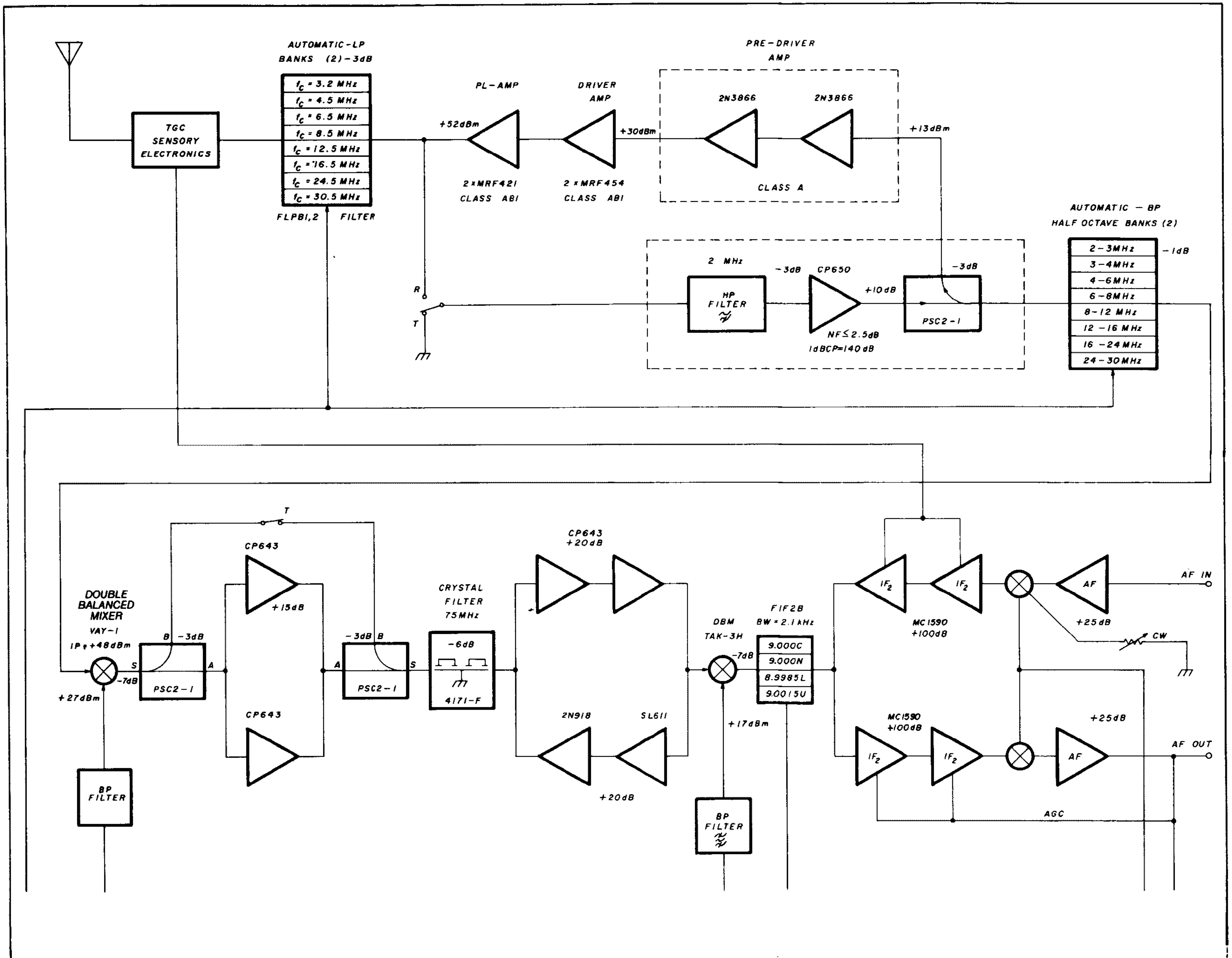

TGC SENSORY ELECTRONICS
AUTOMATIC-LP BANKS (2) -3dB
f_c = 3.2 MHz
f_c = 4.5 MHz
f_c = 6.5 MHz
f_c = 8.5 MHz
f_c = 12.5 MHz
f_c = 16.5 MHz
f_c = 24.5 MHz
f_c = 30.5 MHz
FLPB1,2 FILTER
+52dBm
PL-AMP
2 x MRF421 CLASS AB1
DRIVER AMP
2 x MRF454 CLASS AB1
+30dBm
PRE-DRIVER AMP
2N3866
2N3866
CLASS A
+13dBm
R
T
2 MHz
HP FILTER
-3dB
CP650
+10dB
NF≤2.5dB
1dBCP=140 dB
-3dB
PSC2-1
AUTOMATIC - BP HALF OCTAVE BANKS (2)
2-3MHz
3-4MHz
4-6MHz
6-8MHz
8-12 MHz
12-16 MHz
16-24 MHz
24-30 MHz
-1dB
DOUBLE BALANCED MIXER
VAY-1
IP ≥ +48dBm
+27dBm
-7dB
S
B
-3dB
PSC2-1
A
BP FILTER
T
CP643
+15dB
CP643
-3dB
B
A
S
PSC2-1
CRYSTAL FILTER 75MHz
-6dB
4171-F
CP643 +20dB
2N918
SL611
+20dB
DBM
TAK-3H
-7dB
+17dBm
BP FILTER
FIF28
BW = 2.1 kHz
9.000C
9.000N
8.9985L
9.0015U
IF_2
IF_2
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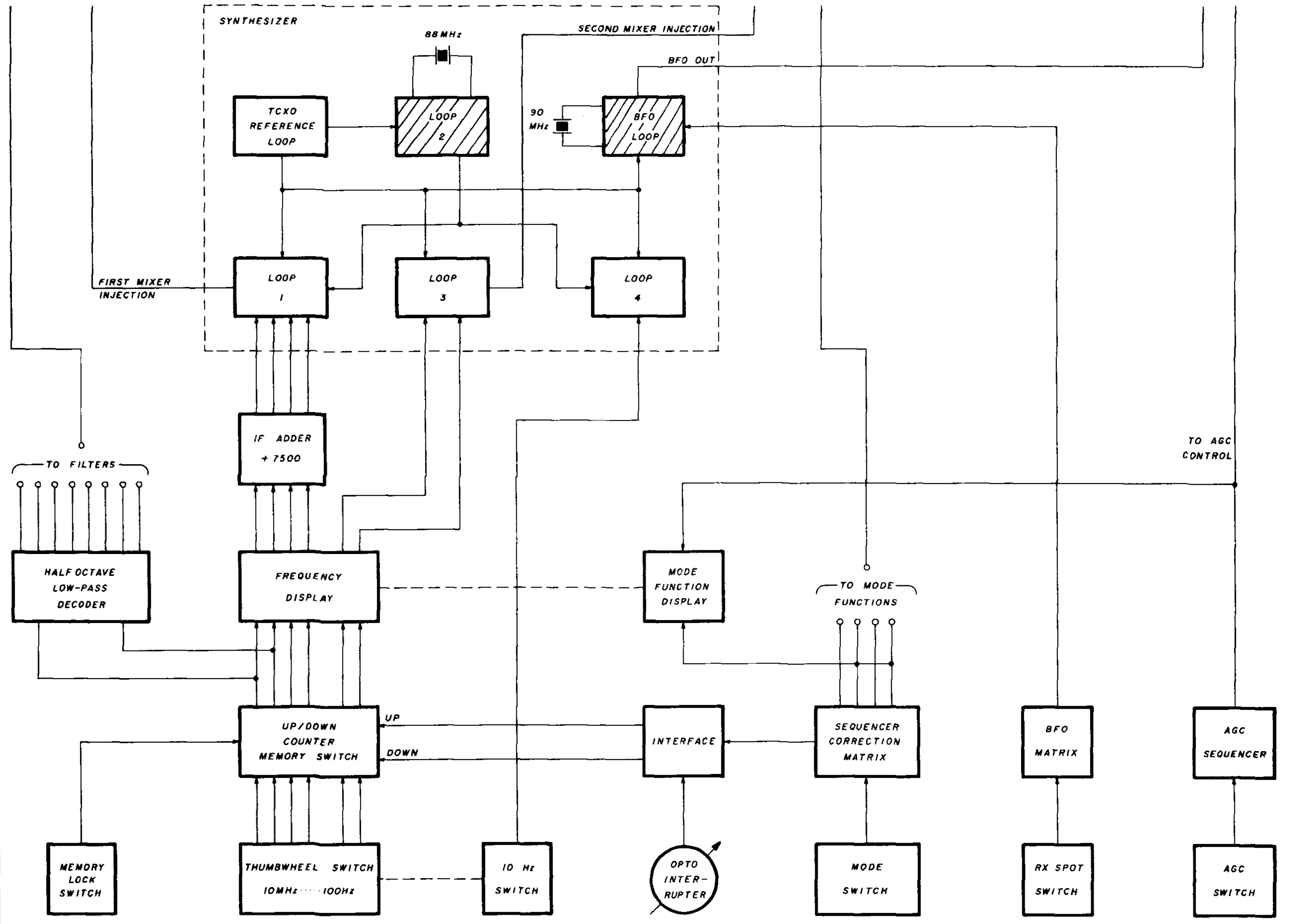


fig. 1. Application of fixed VHF phase-locked-loops in a fully-synthesized HF transceiver. The VHF loop described here is used twice (88 MHz and 90 MHz for the BFO) in order to ensure full synthesis for the entire radio.

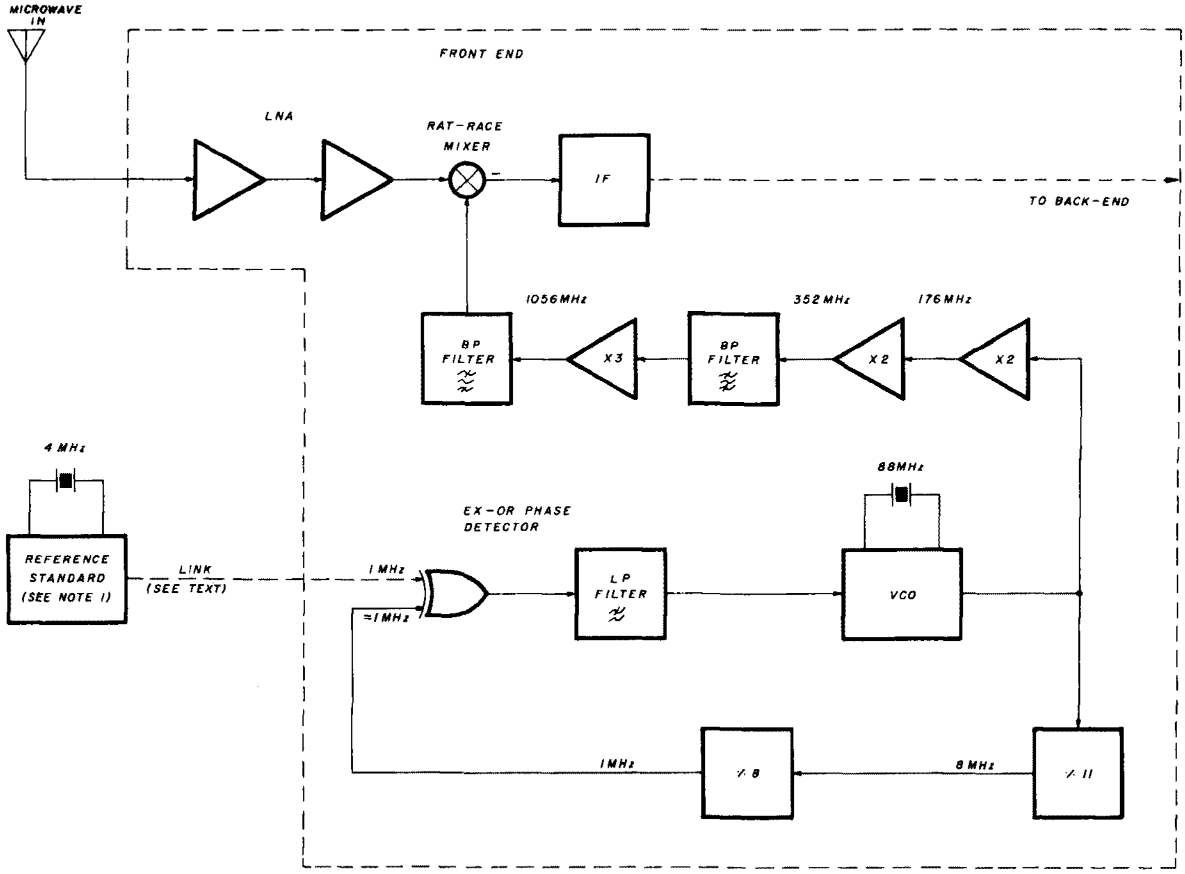


fig. 2. Block diagram of the phase-locked-loop as used in a stable microwave local oscillator located at the low noise block converter (LNB) outdoors. The 1 MHz reference frequency is derived from an indoor TCXO.

used in a series resonant Colpitts oscillator with one side of the crystal grounded. I selected the Colpitts design because of its well known circuit stability and predictability.[5] The output of the oscillator is then amplified and converted to the ECL level required by the 95H90 divide-by-11 device.[6,8]

An ECL-to-TTL level shifter circuit follows the 95H90 and additional division is obtained with three divide-by-2 sections of 74LS74 devices.

The divided-down 1-MHz input to the phase comparator 74L86 (slow L logic was chosen here to keep the output as quiet as possible) is compared against a true 1-MHz reference available from the back end. The exclusive OR phase detector was chosen as a perfect application for this type of phase-locked-loop because the crystal controlled Colpitts oscillator is guaranteed to start almost on frequency and within the capture range of this particular type of phase detector which is only 2π.[2,3,7] Other important criteria for choosing this type of phase detector were the 50 percent duty cycle of the signals present at its inputs and the high reference frequency (1 MHz), all design requirements for successfully using an exclusive OR gate as a phase detector.[7]

When the loop is locked, the output of the phase detector is a 50 percent duty cycle square wave at twice the reference frequency; in this case it is 2 MHz as shown in fig. 4. This output is averaged by the simple RC low-pass filter with a corner frequency (ω_η) of about 16 Hz and is finally presented to the Amperex varactor diode BB-109, which acts as a variable capacitor steering the crystal oscillator in lock. Under locked conditions, the averaged DC output of the phase detector equals half of the TTL supply voltage, or about 2.5 volts DC. When this occurs, the divided 1-MHz signal present at one of the phase detector inputs lags the reference 1-MHz signal present at the other input

*Among many ECL dividers, the 95H90 or its equivalent, the Plessey SP8640 can be programmed to be divide-by-10 or 11, or both in high resolution dual modulus phase-locked-loops.[2]

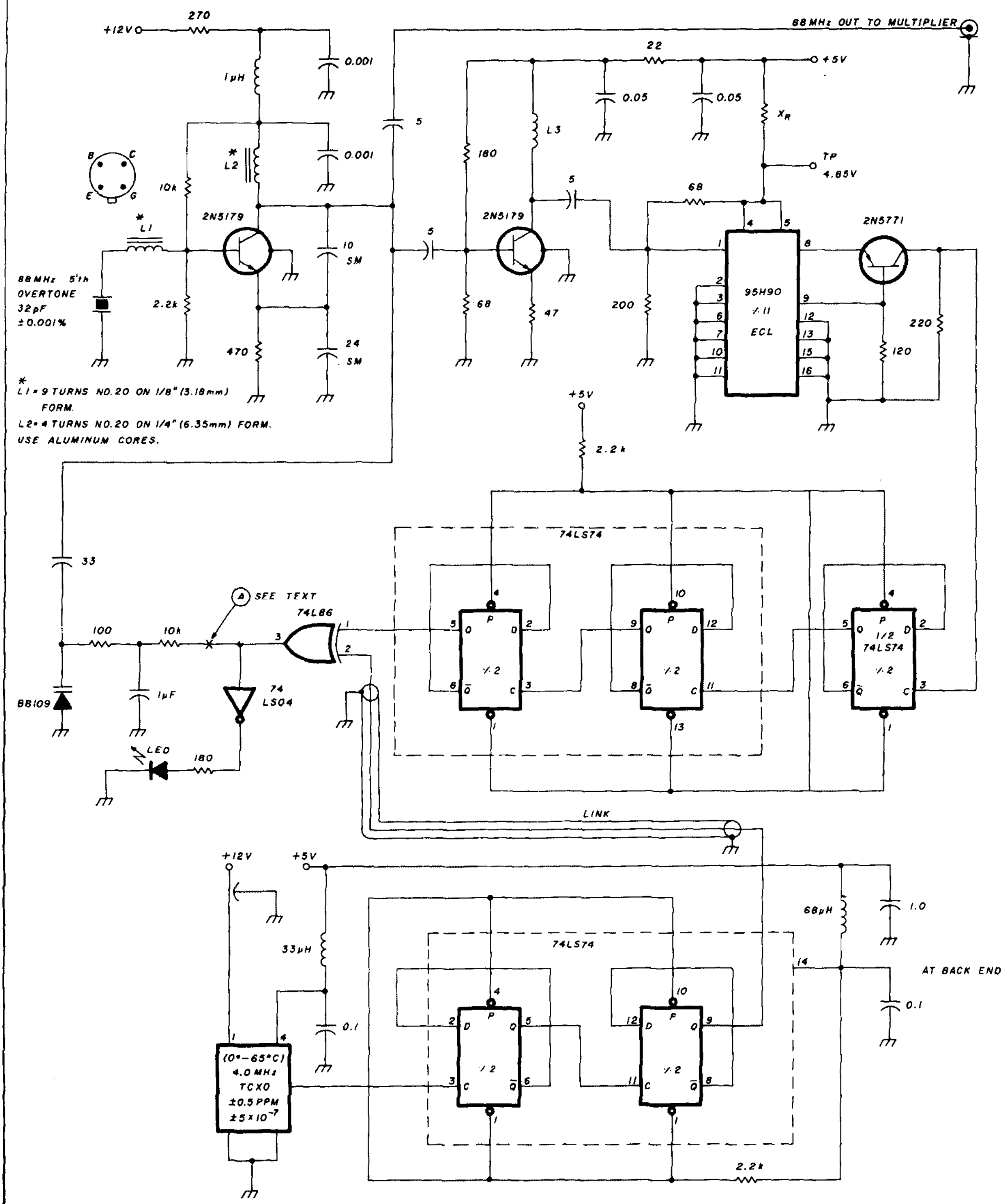


fig. 3. Circuit diagram of the phase-locked-loop. A fifth overtone series-tuned Colpitts oscillator was chosen for stability. The design results in simple and effective exclusive OR phase detectors and low-pass filter. The relatively high 1 MHz reference allows for clean and stable LOs (see text).

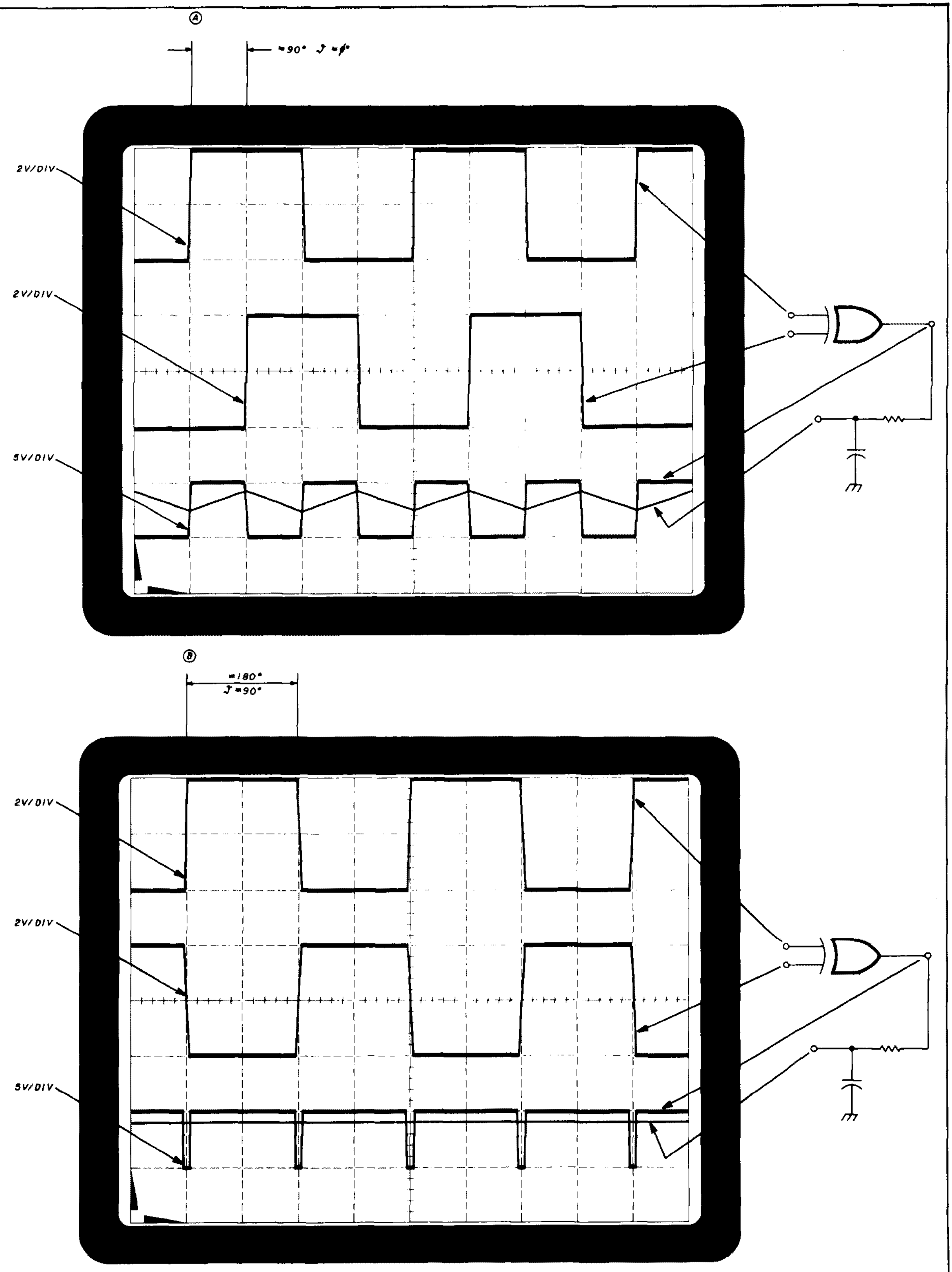
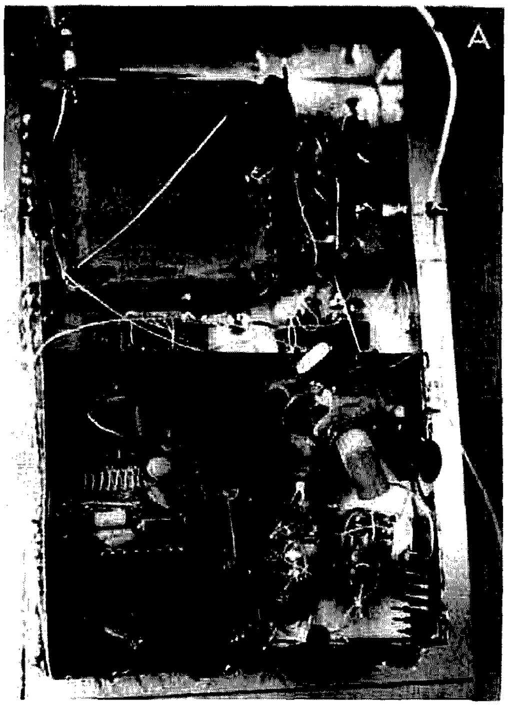


fig. 4. (*A*) Locked condition. (*B*) Out-of-lock conditions for a synthesizer using the exclusive OR phase detector. The output of the low-pass filter is a 2.5 volts DC signal (see text).



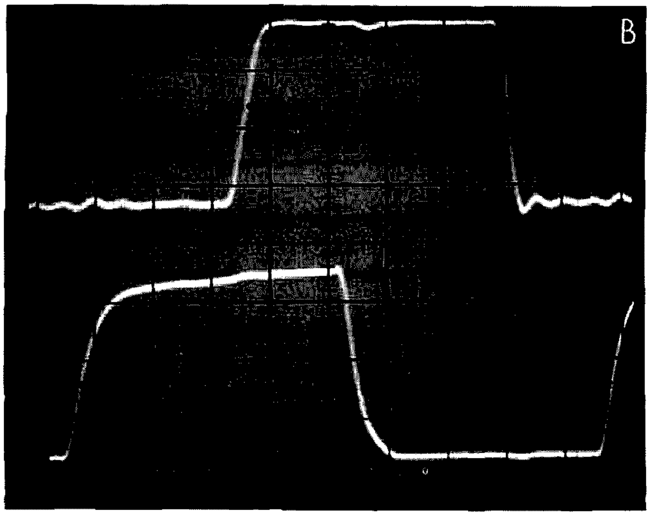


fig. 5. (*A*) The synthesizer circuit was frozen to −78 degrees F with cooling spray. (*B*) After calibration the phase-locked-loop remained locked over the entire temperature range. The divided 1 MHz square wave shown at the top follows the reference frequency at the bottom by −90 degrees, which represents a 0-degree phase error (see text).

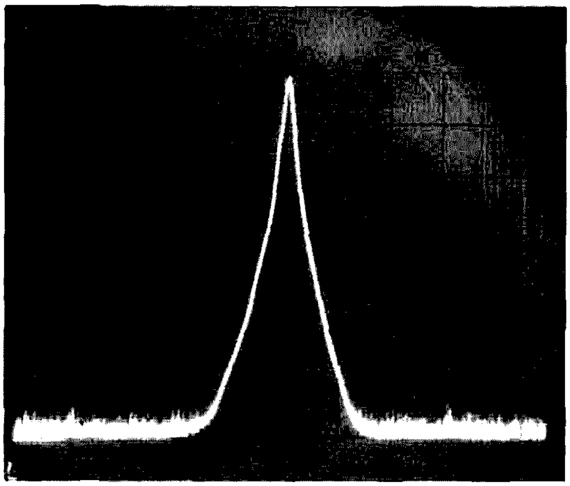


fig. 6. Spectrum analyzer tests performance of the phase-locked-loop.

by approximately 90 degrees (delay is introduced by the additional circuitry) which corresponds to a zero phase error.[3] The secret of the entire circuit is L1 and L2, which are calculated to resonate the Colpitts oscillator on the fifth overtone of the crystal. Additional "tweaking" may be required to bring the circuit into resonance due to local stray elements. High *Q* aluminum cores were used in my prototype for best results.

adjustments

With the loop line disconnected at A (see **fig. 3**), inject a 2.5 VDC level at the 10 kilohm resistor (which is part of the low-pass filter. The best way to do this is to use a couple of batteries in series through a 10-kilohm linear potentiometer voltage divider to insure a pure DC voltage. With the circuit board heated to 80-degrees F (a 100-watt lamp on top of the circuit will do well), adjust L1 and L2 for resonance at the fifth overtone of the crystal as measured on a frequency counter. Observe the position of the cores. Cool the circuit to −78 degrees F with a dry cooling spray. Adjust L1 and L2 again for resonance. (The circuit should still work at this temperature; according to calculations, the transistor junctions will reach only about −40 degrees F.) To accomplish this, selected parts have been used in the prototype. Again, observe the new position of the cores.

Wait until the circuit returns to room temperature and readjust L1 and L2 midway between the two positions. Remove the 2.5 VDC voltage from the low-pass filter and reconnect the loop back at A.

The circuit should now lock every time power is turned on and lock should be maintained over the entire temperature range. This can be verified through repeating the above procedures with the loop closed.

Figures 5A and **B** show the "brassboard" of the synthesizer and how phase locking was maintained under freezing conditions. For convenience a lock indicator was incorporated in the design as shown in **fig. 3**.

An out-of-lock condition would be indicated by the blinking LED, should recalibration become necessary.

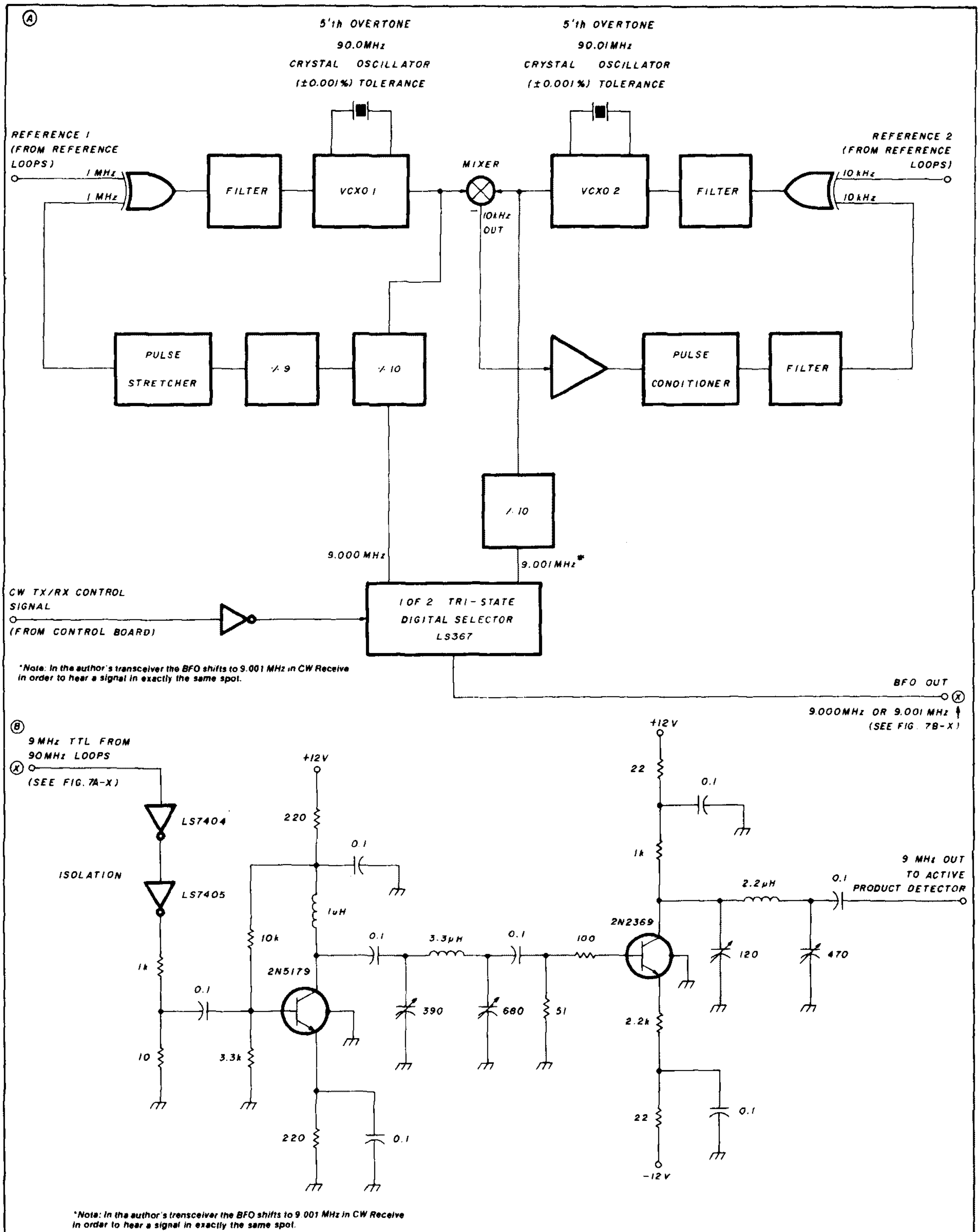


fig. 7. *(A)* **Block diagram of an elegant 9 MHz BFO as used in a fully synthesized HF transceiver. The signal is derived from a dual 90 MHz phase-locked-loop similar to the one presented here (see text).**
(B) **Signal processing for the 9 MHz BFO. The 90 MHz derived signal is conditioned before it is applied to the active mixer product-detector/modulator.**

If the circuit is to be used at room temperature, such as in an HF or VHF/UHF receiver, no indicator should be required. Other status reporting features are possible and can be remotely monitored at the back end of the microwave receiver.

This synthesizer represents an effort of several weeks of design and "brassboarding." It began with the use of a 54S124 dual VCXO integrated circuit that was intended to work at 85 MHz, but was not successful for my application because of the crystal frequency restrictions and the limited upper frequency range for this device.

The new design will work to about 140 MHz (1680-MHz LO output when used with the N6TX multiplier, and beyond with other multipliers). Its limitations are the fabrication of crystals at VHF frequencies and the rather small size of L1 and L2 at the higher frequencies.

This design represents a practical approach to clean and stable local oscillators. Phase and amplitude noise have been measured to be at least −70 dB/Hz at ±100 Hz from the desired carrier (see **fig. 6**). Discrete spurious components were better than −60 dB, while the wide band noise was at least −70 dB measured with a 10-kHz bandwidth.

These specifications depend on the application of sound RF design techniques and may vary according to the circuit components selected. Although a synthesizer layout is not provided in this article, compartmentalization of modules in a true "synthesizer fashion" is highly recommended.

As suggested at the beginning of this article, other than microwave applications of this synthesizer are possible. Among them are beat frequency oscillators (BFO) and fixed oscillators used in multi-loop fully synthesized HF and VHF/UHF receivers and transceivers, as shown in **fig. 1**. **Figure 7A** shows the block diagram of a 9-MHz BFO application as used in my HF transceiver. **Figure 7B** shows the circuit details of the digital to analog portion of the BFO which provides the proper injection to an active product detector/modulator in my transceiver. Many other applications are possible.

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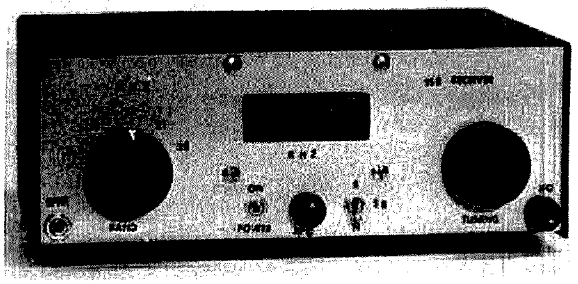

10 through 80-meter homebrew receiver

Even the beginner can build and understand these circuits

If you're looking for a ham band receiver that combines good performance with simplicity of construction and low cost, read on — because this article is for you. This receiver covers the 80, 40, 20, 15, and 10 meter bands and can easily be modified to include the WARC bands or other shortwave segments between 3 and 30 MHz.

This project evolved in response to discussions with other hams about the deplorable decrease in the number of new Amateurs. Traditionally, a large part of our newcomers have come from the ranks of high school students; at one time, young people interested in the electronics almost inevitably gravitated to Amateur Radio because it was easy to get into, with little technical knowledge and only a small investment required. In recent years, the computer hobby with its inexpensive, easily accessible hardware — has provided similar easy entry.

Many hams have expressed the opinion that entry-level equipment has become too expensive and that construction articles now seem to deal with equipment that is complex, expensive to build, and requires a great deal of technical sophistication to construct. This started me to thinking about what it would take to build a receiver for the main ham bands that would provide good performance, yet be simple and inexpensive to build.

the VFO is important

In the 1950's and 60's most receivers used oscillators that switched coils for each band. In later years, however, multiple conversion schemes used crystal oscillators to establish each band and one non-switched oscillator as the variable frequency element to tune within each band. Many receivers now use frequency synthesized oscillators. These latter two concepts, while improving stability, unfortunately increase cost and complexity by a substantial amount.

Recalling more recent commercial rigs, such as the Atlas series, that used switched coil oscillators, I wondered how difficult it would be to design and construct a VFO that had acceptable stability for SSB reception (see **fig. 1**). After deciding it was worth a try, I chose 9 MHz as the IF because of the wide availability of good, inexpensive crystal filters at this frequency. For maximum utility I wanted coverage of both the CW and SSB portions of each band. The VFO covers 500 kHz on 80 through 15 meters and over 1 MHz on 10 meters. As we will see later, these ranges can be easily changed to suit the user's preference.

The choices for the IF and band ranges determine the VFO frequencies as follows:

band meters	VFO frequency MHz
80	12.5-13.0
40	16.0-16.5
20	5.0- 5.5
15	12.0-12.5
10	19.0-20.0

Varactor diode tuning of the VFO was chosen because it helps eliminate hard-to-find mechanical

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variable capacitors as well as the mechanical linkage they require. It also allows size reduction by eliminating the bulk of a variable capacitor and allows significant reduction in the overall mechanical complexity of the VFO unit. One disadvantage of varactors, however, is that they require a tuning voltage as close to pure DC as possible. I therefore decided to use a 9-volt alkaline battery on the prototype. This led to thoughts of running an entire receiver on batteries, in the interest of even greater simplicity.

Since the VFO frequencies for 80 and 15 meters are close, one coil was used for both (see **table 1**). The VFO coils could have been switched electrically with diodes, but doing so would have taken about 30 mA of additional current from the main battery supply. A rotary switch was chosen for current economy.

board construction

The VFO, the frequency counter, the BFO, the bandpass filter, and the IF unit are all constructed in a similar fashion. With the method used here you can actually build any unit in the same or less time than it takes to design and etch a printed circuit board. (One main advantage of etched boards, however, is saving time on successive boards — but since I seldom build more than one of any particular project, PC boards are not time-effective for me.)

I use epoxy material punched with holes on a 0.1-

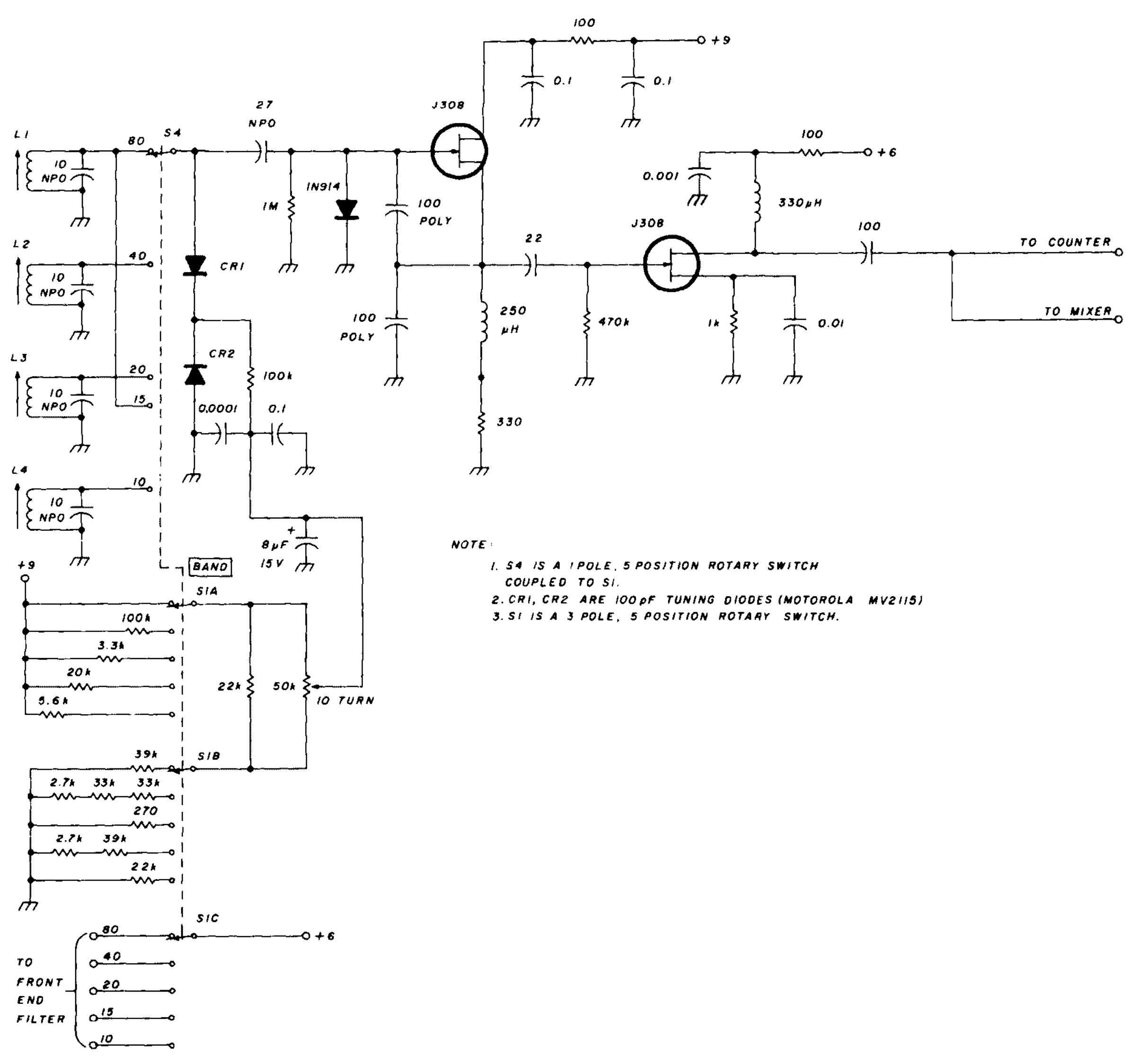
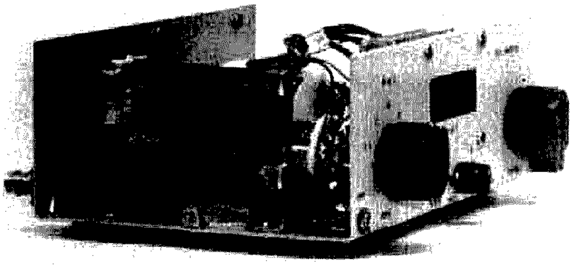


fig. 1. VFO schematic.

table 1. VFO coil data.

coil number	band	turns on 1/4" form*	frequency range (MHz)	wire No.
L1	80	21	12.5-13.0	26
L2	40	15	16.0-16.5	26
L3	20	48	5.0- 5.5	32
L1	15		12.0-12.5	
L4	10	10	19.0-20.0	22

*Coil forms Millen 69043 used but any good 1/4-inch (0.635 cm) ceramic form should be suitable.



Left side view shows bandswitch and ganged VFO switch. Vertical board in center is BFO; VFO board and coils are mounted vertically against rear panel.

inch grid and clad with copper on one side, widely available from mail order suppliers.* Insulated circular pads for IC pins and other components are cut with a pad cutter tool,* either by hand or with the tool in a drill.

For the VFO I disassembled a single-pole, six-position rotary switch and mounted a 3-inch (7.6 cm) square piece of the epoxy material between the switch wafer itself and the switch index mechanism. Three holes were drilled in the material: one for the switch shaft and two that coincide with the wafer mounting screws on the switch itself. (You may have to obtain mounting screws a little longer than those that come with the switch to allow for the 1/16 inch (0.16 cm) thickness of the copper clad board.)

Next cut appropriate pads for the coils and the VFO circuitry. Use T-44 push-in pins,* for terminals such as transistor connections and all input and output connections. The four coils are mounted on the front (non copper-clad) side of the board by passing their leads through and soldering to the pads on the other side which directly connect to the switch. The coil bodies are then covered with clear Silicone II, an adhesive that bonds them firmly to the board and prevents any movement of either the coil bodies or the windings on the coils. This is extremely important in that it is a significant factor in the frequency stability of the VFO. All wiring on the VFO should be direct and rigid, especially the jumper on the switch for the 80-meter position. Sound VHF wiring practices should be followed.

VFO adjustment

After wiring is completed, a test jig should be assembled as indicated to determine the values of the series voltage-dropping resistors necessary to have the tuning control, a 10 turn, 50 kilohm potentiometer, cover the desired range of frequencies (see **fig. 2**).

VFO adjustment proceeds as follows:

- Connect a frequency counter to the VFO output.

*All items marked with an asterisk are manufactured by Vector, Inc.

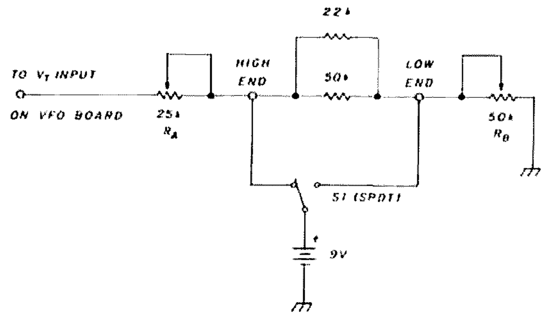


fig. 2. Test jig for determining the tuning voltage divider series resistors for each band.

- Set the VFO to the 10 meter position.
- Set S1 on the test jig to the *HI* position (at end of potentiometer R_A).
- Adjust R_A so that the VFO frequency is about 10 kHz above the upper limit wanted – for example, on 80 meters set R_A so that the VFO output is 13.010 MHz.
- Put S1 in the *LO* position at the end of R_B.
- Adjust R_B until the VFO output is about 10 kHz below the lower limit wanted – for example, on 80 meters set R_B so that the VFO output is 12.49 MHz.
- Put S1 back to the *HI* position and recheck the frequency. Since these settings interact it may be necessary to repeat the procedure. Measure the values of R_A and R_B and use fixed resistors of those values on the bandswitch.
- Repeat this procedure for the remaining band positions on the VFO. It may be necessary to adjust the coil slugs to obtain the required range. On 40 meters it may be necessary to place an approximately 68 kilohm resistor in series with the 9-volt source and R_A.

Because it's unlikely that your unit will require the

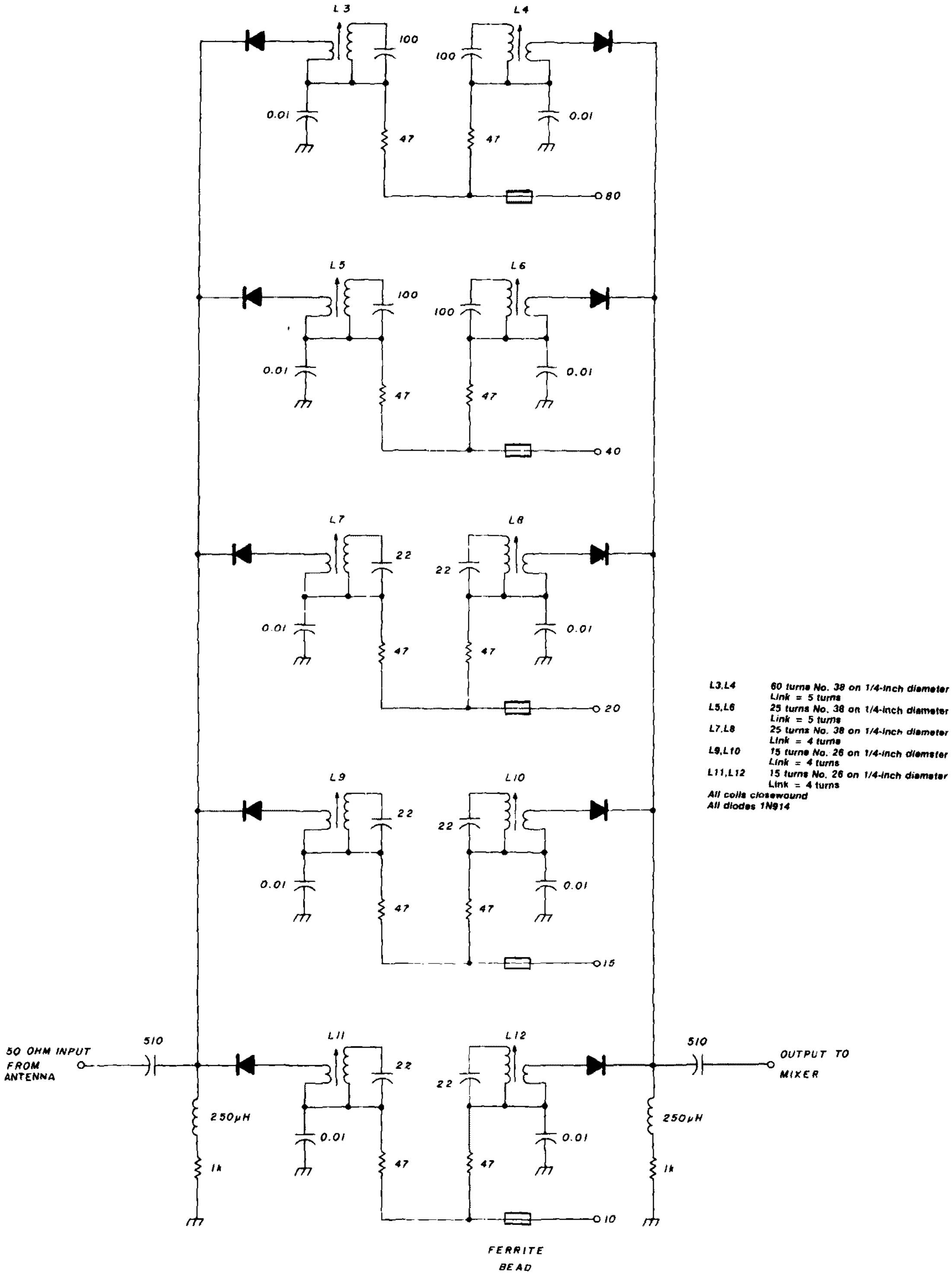


fig. 3. Bandpass filter schematic.

same values of series resistors as mine because of differences that will occur in many components, I advise going through the procedure. I also recommend using alkaline batteries for both 6 and 9-volt supplies. Because of other connections to the 9-volt supply, six AA cells are used; the 6-volt supply consists of four D cells.

bandpass filters

These components are enclosed in a separate unit made entirely of copper-clad punched board. The two coils for each band are placed next to each other on 1/4-inch (0.635 cm) centers. Six volts applied to a band terminal forward biases the respective diodes and allows signals from the antenna to pass through the resonant circuit. The use of diodes, while drawing additional current, significantly reduces the mechanical problems associated with rotary switches.

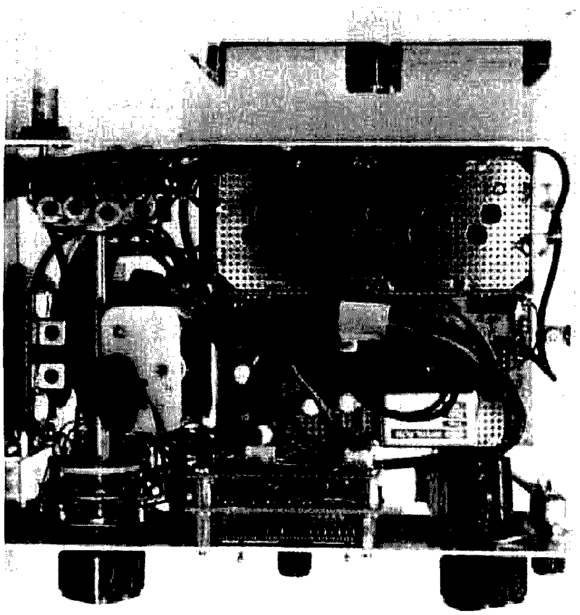
Adjustment proceeds in the following order:

- Connect an antenna to the input and a receiver to the output.
- Set the receiver to 80 meters.
- Apply 6 volts DC to the 80-meter terminal.
- Set the receiver frequency 1/4 of the way up from the bottom of the band, at approximately 3.625 MHz.
- Adjust one of the 80-meter bandpass coils for maximum received signal (or noise).
- Move the receiver to a frequency 1/4 of the way down from the top end of the band, at approximately 3.875 MHz.
- Adjust the other 80-meter bandpass coil for maximum response.
- Repeat as necessary to compensate for interaction.
- Do the same for the remaining four bands.

These settings will establish filter response across the entire bands. If you're primarily interested in either SSB or CW, the coils can be peaked accordingly. If you don't have a receiver or cannot borrow one, it's possible to peak the bandpass filters with the receiver itself, once completed.

BFO

This unit is constructed on a separate board but could be placed on the IF board. Using an available crystal filter translated to a USB BFO frequency of 9.000 MHz and an LSB BFO frequency of 9.0030 MHz (see **fig. 3**). Normal (i.e., non-inverted) sideband in this receiver uses the 9.000 MHz BFO crystal, so no readout offset is necessary. In the reverse sideband position the readout will be off by 3 kHz. The BFO circuit uses series resonant crystals. L1 and L2 are 10.7 MHz IF transformer primary windings (4 μH) with the



Top view shows bandpass filter (top left), IF strip (top center); VFO board (bottom left), BFO board (bottom center); and counter boards (right center). 6-volt battery pack is mounted at rear.

resonating capacitor removed. The 260 and 220 pF capacitors are polystyrene and the 100 pF capacitors are silver mica. To adjust, connect a counter to the output and apply 6 volts DC to the BFO oscillator desired. Adjust the corresponding series inductance for the correct output frequency.

IF board

This board contains the double-balanced mixer, crystal filter, IF amplifier, and audio stages (see **fig. 4**). The only adjustment here is to peak the slug in T1 for maximum signal. T1 is also a 10.7 MHz IF transformer with an external capacitor to bring it down to 9.0 MHz. RF derived AGC is internally generated in the Plessey SL6700 IC. U4 is a Texas Instruments TL442 double-balanced mixer. Connection of pins 3 and 12 result in a 500-ohm output impedance to match the crystal filter. If your filter has a different input impedance you may have to design an appropriate matching network. Application of 9 volts to the mixer is necessary to achieve adequate gain. U6 is an audio preamp and U7 provides up to a quarter watt of audio output to an 8-ohm speaker or headphones. The SL6700 has a maximum voltage rating of 7 volts.

frequency counter

This unit uses LCD and CMOS components because both draw little current and also because we need to count frequencies up to 20 MHz (see **fig. 5**). The least

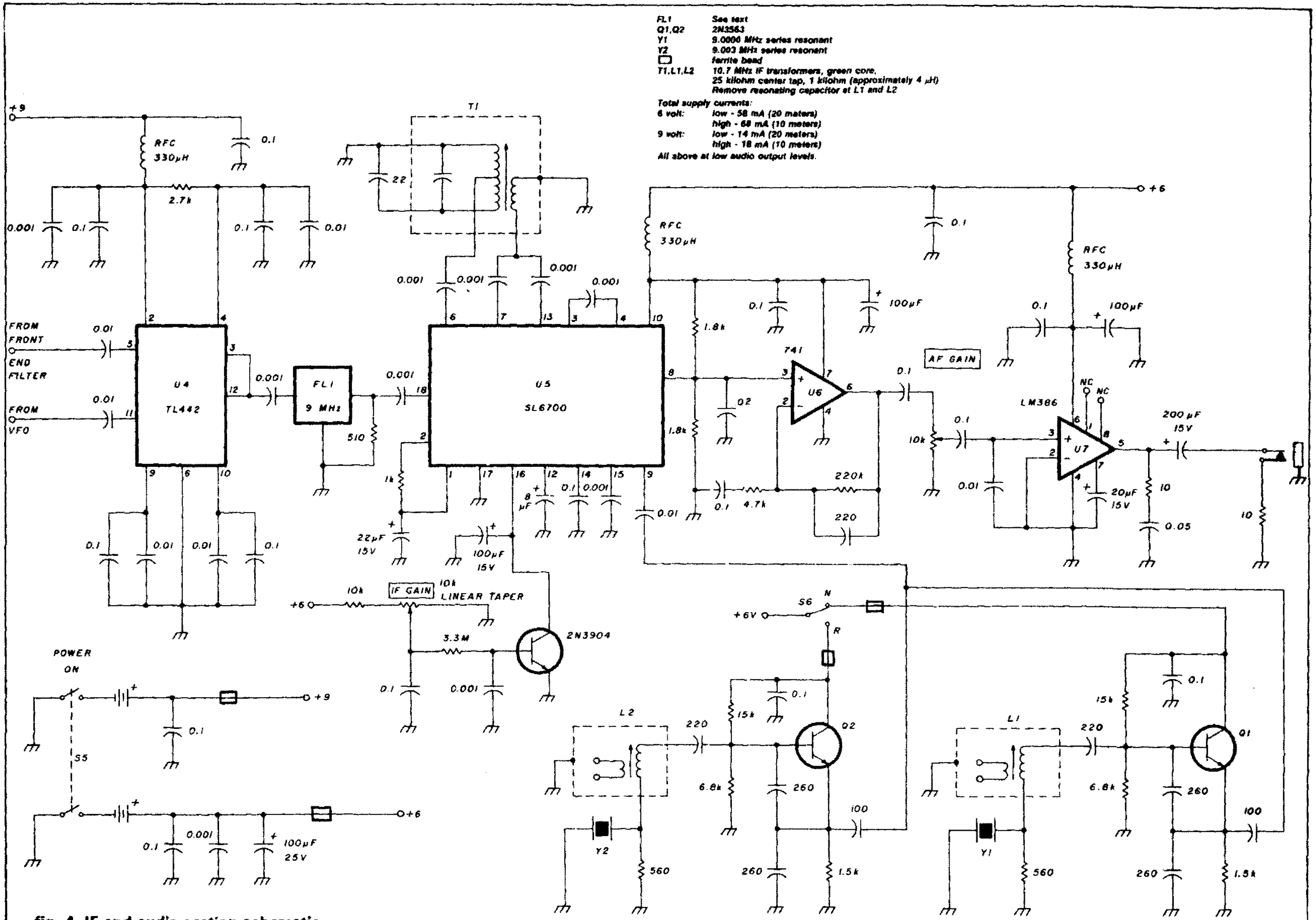


fig. 4. IF and audio section schematic.

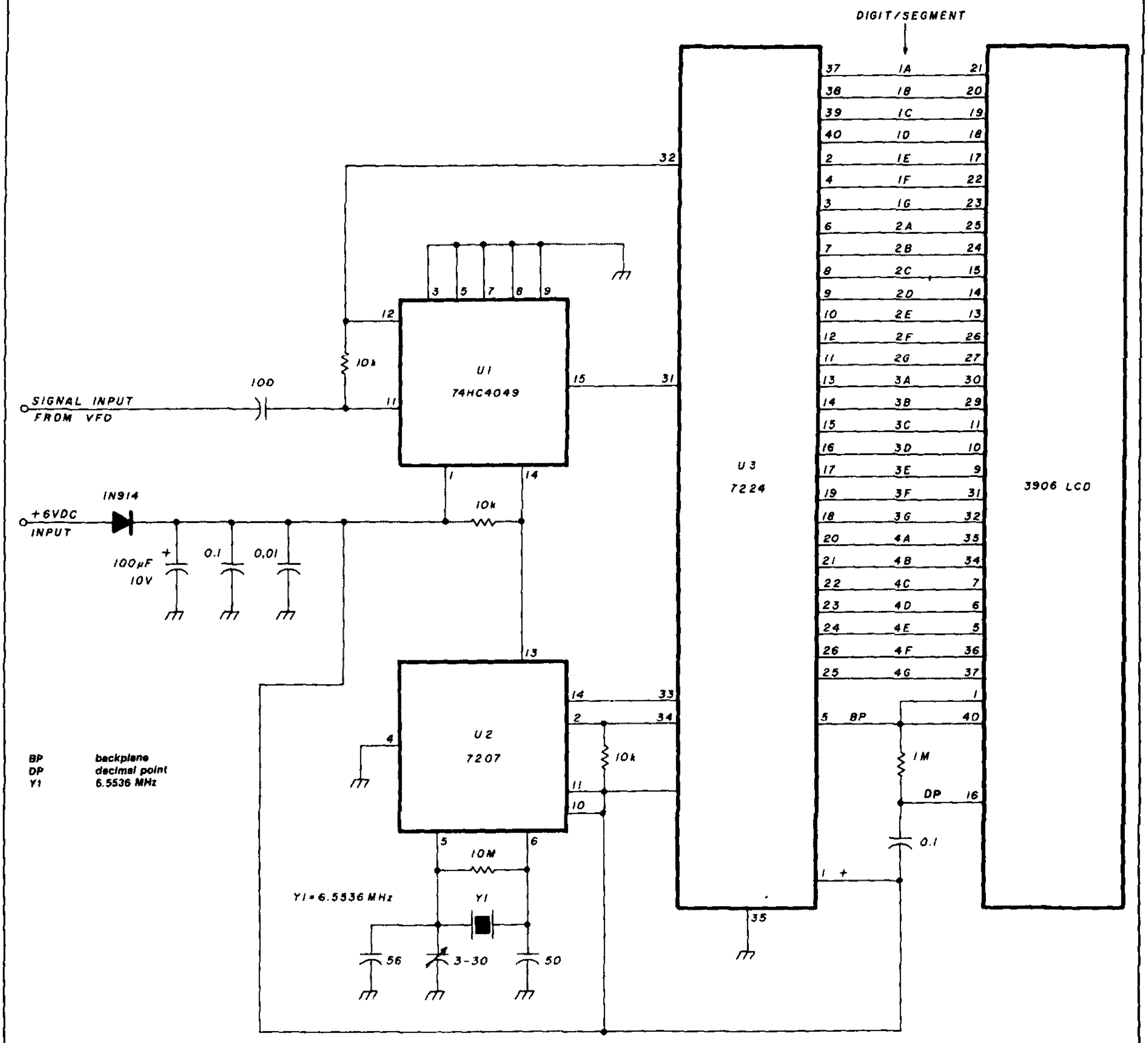


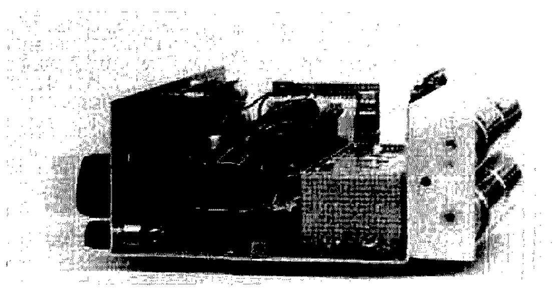
fig. 5. Frequency counter schematic.

significant digit reads out in hundreds of Hertz because of a 10 millisecond gate provided by U2 and its associated crystal. This, and the counter chip, U3, are Intersil ICs. (It's important to note that Intersil also manufactures a 7207 ***A***. It's only through the use of the 7207 — without the "A" — and the specified crystal frequency that the 10 millisecond gate time can be obtained, so be careful when ordering.)

Note that U1 is a high-speed version of the CMOS type 4049. Because the latter type will not function up to 20 MHz, the 74HC4049 is used instead.

The liquid crystal display is a Hamlin type 3906. It's a little tricky to identify pin 1; to do so just hold the display in front of you and move it so that light reflects off its face. You'll notice the outlines of the 8s that are the digits. Look for the three decimal points between the 8s. Rotate the display if necessary so that the decimal points are at the bottom. The left end pin on the top row is pin 1. These both connect to the display backplane (see **table** 2).

The counter is constructed on two pieces of perforated board 2-1/2 inches (6.35 cm) square, separated from each other and the receiver front panel by 1/2 inch (1.27 cm) threaded metal standoffs at each corner. The socket for the 7224 IC is a low profile type and is nested inside the socket strips for the LCD. These strips are merely the halves of a 40-pin IC socket separated by cutting it lengthwise with a hacksaw.



Right side view shows tuning pot and IF gain pot.

table 2. 3906 and 7224 designations.

3906 pins				7224 pins			
20	1B	1A	21	20	4A	4B	21
19	1C	1F	22	19	3F	4C	22
18	1D	1G	23	18	3G	4D	23
17	1E	2B	24	17	3E	4E	24
16	DP *	2A	25	16	3D	4G	25
15	2C	2F	26	15	3C	4F	26
14	2D	2G	27	14	3B	1/2 digit	27
13	2E		28	13	3A	CO	28
12		3B	29	12	2F	LZ1	29
11	3C	3A	30	11	2G	LZO	30
10	3D	3F	31	10	2E	C	31
9	3E	3G	32	9	2D	inhibit count in	32
8	4DP		33	8	2C	RST	33
7	4C	4B	34	7	2B	ST	34
6	4D	4A	35	6	2A	ground	35
5	4E	4F	36	5	BP	OSC	36
4		4G	37	4	1F	1A	37
3			38	3	1G	1B	38
2			39	2	1E	1C	39
1	BP	BP	40	1	+	1D	40

*Use this decimal point

These two sockets are mounted on unclad perfboard. The rear board is copper clad and contains the signal processing and gating ICs, U1, and U2. Because of the relatively high frequencies involved, this hoard *must* be copper clad.

Upon completion of the counter wiring connect a clip lead between pin 29 on U3 and ground. For test purposes this will cause all digits to appear as zeros with no signal input. Connect the 6 volts to the counter; application of voltage greater than 6.5 to U3 could destroy the IC. The purpose of the 1N914 in the counter is to reduce the 6 volts slightly. Four zeros should appear on the display with a decimal point between the least significant zero and the next one.

To calibrate the counter, run a short length of RG-174/U coax from the output of the BFO to the counter input. Connect an external frequency counter to the same point. Adjust the 30 pF trimmer capacitor at U2 until the last four digits of the two counters are the same. Remove the clip lead from the 7224. Leading zeros will now be blanked on the display.

general construction notes

The receiver is built in an aluminum enclosure available from Radio Shack (see table 3). A slightly larger box would have allowed placement of the batteries inside. The band switch is a three-pole, six-position rotary mounted on the front panel. A coupling joins the end of its shaft to the VFO switch shaft. The 9-volt battery pack is attached to the chassis bottom with self-adhesive Velcro strips from Radio Shack. This sturdy arrangement allows easy removal for battery replacement. Four D cells in a holder mounted on the outside rear wall of the receiver supply 6 volts. A jack on the front panel provides audio for external speaker or headphones. All circuits and functional modules are extensively decoupled with ferrite beads and capacitive bypassing. The receiver has no birdies and the decoupling is probably a significant factor. Small diameter RG-174/U coax is used for all signal interconnections as well as to and from the volume control. The only totally shielded unit is the bandpass filter unit. The IF gain control carries DC and does not require shielding.

changes

Few designs remain unchanged for long. Were I to build this one again, I would consider making some

table 3. Parts and source list.

part	source
Hamlin 3906 LCD, MV2115 diodes, 74HC4049, J308, 741, LM386, 10.7 MHz IF transformers, Vector board	Circuit Specialists PO Box 3047 Scottsdale, Arizona 85267
Intersil 7224 counter IC	Jameco Electronics 1355 Shoreway Road Belmont, California 94002
6.5536 MHz crystal	Digi-Key Corp. 701 Brooks Avenue South Thief River Falls, Minnesota 56701
Intersil ICM7207 IC	Advanced Computer Products PO Box 17329 Irvine, California 92713-7329
Texas Instruments TL442 IC	Radiokit PO Box 411 Greenville, New Hampshire 03048
Plessey SL6700 IC	Circuit Board Specialists PO Box 969 Pueblo, Colorado 81002
BFO crystals	JAN Crystals 2400 Crystal Drive PO Box 06017 Fort Myers, Florida 33906-6017
Enclosure, battery holders, miscellaneous parts	Radio Shack

changes. First, this receiver is designed with VFO frequencies chosen so that, for conventional SSB practices, a 9 MHz USB filter and one BFO oscillator and crystal would probably be adequate for most users. Given the current drawn from the D cells, the batteries will probably last their shelf life; it may be possible to change to C cells to save weight and volume. Perhaps the two sets of batteries could be changed to one set of eight C cells with a voltage regulator for the 9 volts and a direct tap for the 6 volts. One change I would definitely make would be separate 80 and 15 meter VFO coils. This would make it much easier to set the required frequency range on both bands.

Other than the 10 pF NPO capacitors across each VFO coil, no further temperature compensation was done since stability at that point was adequate for my purposes. VFO stability could be further enhanced, especially at the higher VFO frequencies. I would also build the VFO in a separate, shielded enclosure, because there is a slight frequency shift when the receiver cover is removed. This small portable receiver meets all my original goals: simplicity, low cost, and good performance. These are exactly the characteristics which make this receiver ideal for newcomer and old-timer alike.

I'll be happy to respond to any questions accompanied by an SASE. (Send to Author's address —Ed.).

ham radio

VHF/UHF WORLD

Joe Reisert W1JR

high dynamic range on 2 meters

In last November's VHF/UHF World column[1] we discussed high dynamic-range (HDR) with emphasis on circuitry for the 6-meter band. Judging from the correspondence I've received and the quantity of parts ordered from Proto Parts,* the interest in this subject is quite high. There were also plenty of inquiries for information about a companion 2-meter down-converter.

Further work has been conducted since reference 1 was published. Much of this new work, which primarily dealt with *antenna pattern* improvements, local oscillator phase noise studies and problems with the IF or receiver, was presented at the Dayton Hamvention.[2]

*Proto-Parts (formerly Proto Fab), 74 Wedgemere Drive, Lowell, Massachusetts 01852.

Since time is precious, I'll keep this month's column short and to the point, concentrating almost entirely on the design of a 2-meter HDR receiver down-converter as shown in block form in **fig. 1**. The material in reference 1 will not be repeated except as required. Further information on this very interesting subject will be published as time permits.

preamplifiers

Let's start by assuming that you've read references 1 and 2. Let's go directly to the preamplifier stage. Common semiconductor devices used on 2-meter preamplifiers include bipolar transistors, JFETs, dual-gate MOS FETs and GaAs FETs. All of these can attain excellent noise figures (less than 2 dB) and more than sufficient gain (12 to 30 dB!).

High gain is not too compatible with high dynamic range since the higher the gain, the more likelihood of compression, overload, and IMD (intermodulation distortion). Furthermore, overdriving the mixer, which results in a lower dynamic range, can result from too much gain in the stage preceeding the mixer.[1]

The HDR *transformer-coupled* lossless feedback preamplifier described in reference 1 can be used as shown to cover the entire frequency range from 1.5 to 200 MHz with almost no loss of dynamic range and only a slight increase (0.5 to 1.0 dB) in noise figure at the higher end of this frequency range. Hence this circuit is highly recommended for a 2-meter HDR preamplifier. However, it requires proper input and output filtering similar to the scheme shown on the 6-meter converter to prevent responding to the entire HF/VHF spectrum! (More on this shortly.)

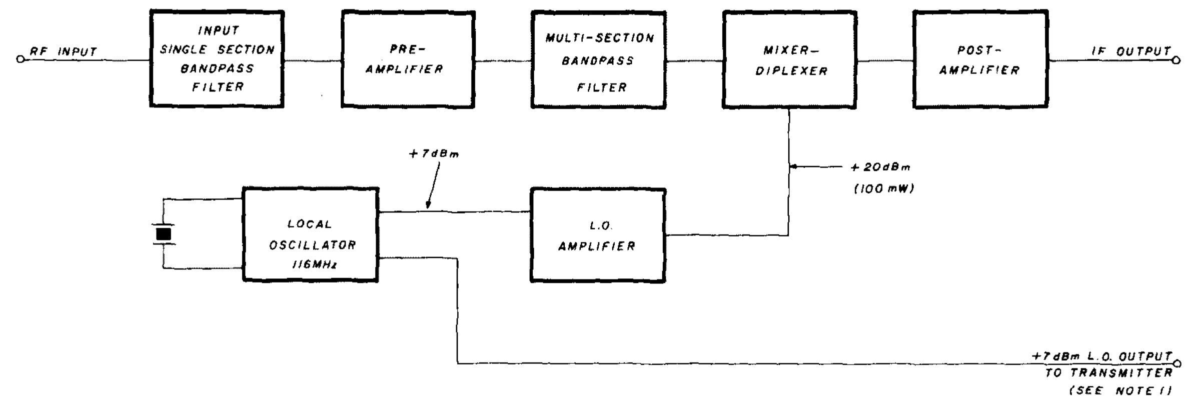


fig. 1. Overall block diagram of a 2-meter HDR receive-type down converter. ***Note 1*****: If the transmitter LO port is unused, terminate it with a 50-ohm load.**

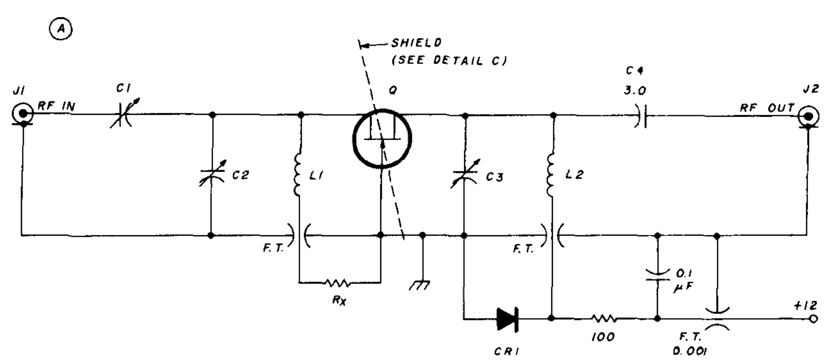


fig. 2A. Schematic diagram for a recommended JFET 2-meter preamplifier with moderate dynamic range. Gain is 11 dB and noise figure is 1.5 to 2 dB typical.

C1,C2,C3	1-10 pF low-loss air variables
CR1	1N4001 or equivalent "idiot diode"
FT	0.001 μF feedthrough type capacitor
J1,J2	BNC connectors
L1,L2	5 turns of No. 18 AWG on 0.25 inch (6.35 mm) ID and 0.5 inch (12.7 mm) long
Q1	U310 JFET (see text)
Rx	Set for I_D of 10 ± 2 mA (150 ohms typical)

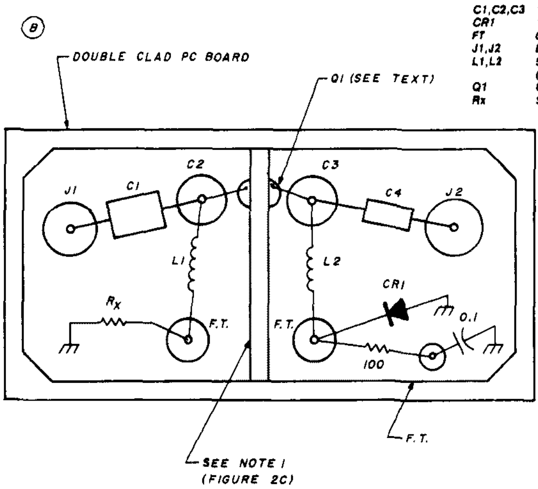


fig. 2B. Bottom view of preamplifier showing recommended construction technique. The preamplifier can be built on the cover of a Pomona 2417, Bud CU123, or equivalent cast box. (See text for other details.)

Dual-gate MOSFETs — and GaAs FETs as well — are not recommended for HDR since they have too much gain. Both are excellent for very low noise figures. The output compression point for GaAs FETS is poor for HDR (+6 to +8 dBm or 4 to 6.3 milliwatts).[3] If you use them they should have built-in bypass relays so they can be switched out of the circuit if there are strong signals present that limit the dynamic range.

One of my favorite quick-and-dirty preamplifiers uses a JFET. I particularly like the U310, a low-cost, readily available JFET. *Caution: don't use the J310, a plastic TO-92 version of this device, above 100 MHz. The plastic package has too much internal inductance in the gate lead and the device will be potentially unstable at 2 meters and above.*

The U310 can deliver good performance from 50-450 MHz with a typical gain of 11 to 13 dB, a reasonable noise figure (1.5 to 3 dB), and very good dynamic range. The typical output compression point is not as high as the HDR lossless feedback circuit and is typically in the range of +10 to +14 dBm (10 to 25 milliwatts), which is more than sufficient for most applications.

A JFET preamplifier offers another advantage in that it has some "built-in" selectivity from its associated matching circuits so it can be operated without additional input or output filtering. Two of these amplifier circuits in cascade will usually have more than adequate selectivity to do the job as is. The circuit I use has been around many years and was discussed in reference 4. The recommended circuit and values for 2 meters are shown in **fig. 2**.

If you don't want to build all the extra filtering associated with the bipolar lossless feedback circuit and can sacrifice a small amount of dynamic range, you can use the JFET circuit shown in **fig. 2**. If only a single stage is used, the overall noise figure of the converter will be approximately 4 dB, hardly a problem for most types of operation. This sure beats many commercial transceivers that usually have a 7-10 dB or higher noise figure along with poor dynamic range and phase noise to boot! Two JFET preamplifier stages can be cascaded for moderate

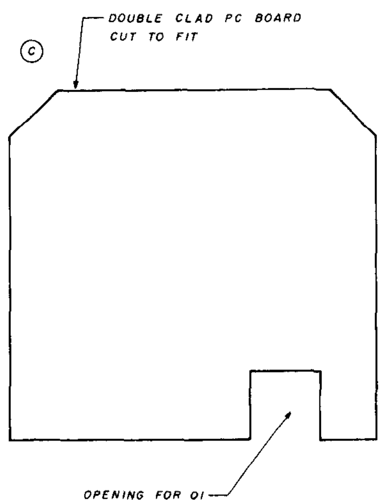


fig. 2C. Typical shield to be placed across Q1. ***Note 1*****: Solder both sides of shield to the ground plane where shown.**

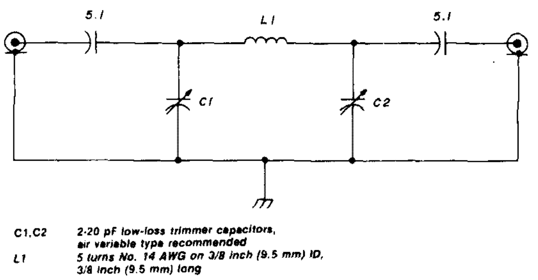


fig. 3. Typical recommended input band-pass filter.

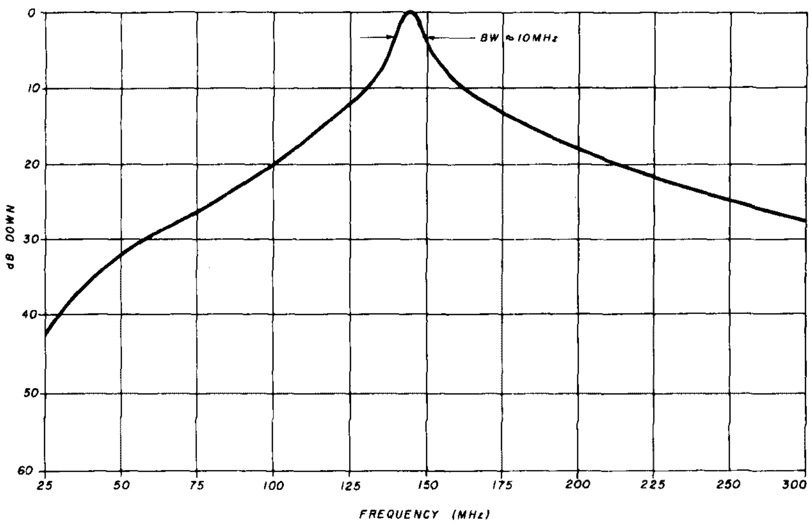


fig. 4. Frequency response of the simple band-pass filter shown in *fig. 3*.

noise figure (1.5-2 dB) with a commensurate loss in dynamic range.

input filters

If the HDR lossless feedback preamplifier is used, input and output filtering are required. A simple single section bandpass should be used on the input for reasons discussed in reference 1. Typically speaking, the capacitance-coupled circuit in reference 1 has a sort of high-pass type of bandpass response,[5] so it is not as effective with all the many types of high power RF emitters that are active on both sides of the 2-meter band.

Figure 3 shows an interesting filter topology.[6] At first glance it looks like a Pi-network. It has one more component than the usual filter, a second tuning capacitor.

The advantage of this filter topology is that it has a symmetrical response about the center frequency with only a moderate loss, 0.3 to 0.5 dB typical. As shown in **fig. 4**, the half-power bandwidth is 10 MHz, while the 10 dB and 20 dB down bandwidths are about 30 and 110 MHz, respectively.

band-pass filters

Again, if a broadband preamplifier is used, additional filtering will be required just ahead of the mixer as discussed in reference 1. A recommended three-section band-pass filter is shown in **fig. 5**. A plot of its filter characteristics is shown in **fig. 6**. It has a fairly symmetrical passband shape and only a moderate loss — 1.3 to 1.5 dB. Typical half-power bandwidth is 10 MHz, and the 10 and 30 dB down bandwidths are typically 15 and 33 MHz, respectively.

The loss in this band-pass filter will not degrade noise figure significantly if it is preceded by at least 9 to 10 dB of preamplifier gain. This filter should provide more than adequate selectivity for a high performance 2-meter receiving converter.

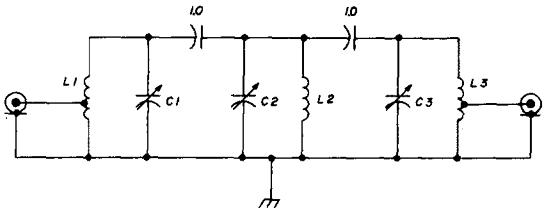
mixers

Various types of mixers were discussed in reference 1. The double balanced mixer (DBM) is highly recommended. The MiniCircuits Labs TAK-1H or equivalent is highly recommended for 2 meters. Other mixers, shown in **table** 1 of reference 1, are also acceptable. The lower drive DBMs such as the SRA-1 will reduce dynamic range but still may be more than sufficient for most applications.

The DBM circuit in reference 1 is very wideband and already covers frequencies through 500 MHz so it will not be duplicated here. *Don't leave out the diplexer*; it's *very* necessary. Also, since the 2-meter band is much higher in frequency, the use of a 14-MHz IF is discouraged since it won't offer sufficient image rejection without extensive filtering.

local oscillator

A Colpitts series-mode oscillator similar to the one discussed in references 1 and 7 is highly recommended. Since the typical IF for a 2-meter converter is 28-30 MHz, a 116-MHz local oscillator is recommended. *Don't use a 58-MHz oscillator with a frequency doubler; it will make breakthrough from TV and FM stations a potential problem. Don't be "penny wise and pound foolish." Remember — this is*



C1,C2,C3	2-20 pF trimmer capacitors
L1,L3	4 turns No. 20 AWG 0.25 inch (6.35 mm) ID and 0.5 inch (12.7 mm) long. Tap is placed 1 turn up from ground end
L2	Same as L1 without tap

fig. 5. Typical recommended three-section 2-meter band-pass filter.

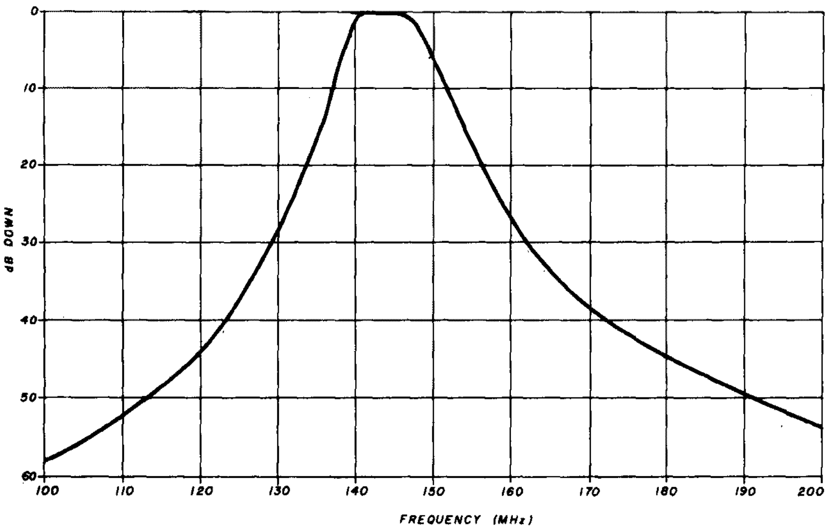


fig. 6. Plot of the frequency response of the three-section 2-meter band-pass filter shown in *fig. 5*.

a high-performance HDR receiving converter!

Since the frequency of the oscillator is much higher than in the 6-meter converter, the tuned components had to be modified. The final oscillator circuit and a suitable power amplifier are shown in figs. 7 and 8. It will provide the necessary 100 milliwatts of power. (More on this later.) Again, a low-level output is provided for a transmitting up-converter.[8]

postamplifiers

The low-noise postamplifier using the lossless feedback circuit described in reference 1 is recommended without changes, so it won't be duplicated here. Note how the modular approach discussed in references 1 and 7 has paid off. If you have the mixer and post-amplifier from the 6-meter converter, all you have to do is to change the local oscillator and front end, and away you go on 2 meters.

construction tips

Once again, the modular approach is recommended, with each separate circuit packaged in a shielded box. The circuits can be built above a double-clad printed circuit type of material as described in reference 1. The boards can be attached to the cover of the boxes with the connectors. This method provides excellent grounding, and the components can be soldered directly to the board where grounding is required.

The three-section band-pass filter must be carefully laid out so that mutual coupling will not cause pass-band ripple. This can be accomplished *if the individual coils are all wound in the same direction and placed at least 1/2 inch (12.7 mm) apart.*

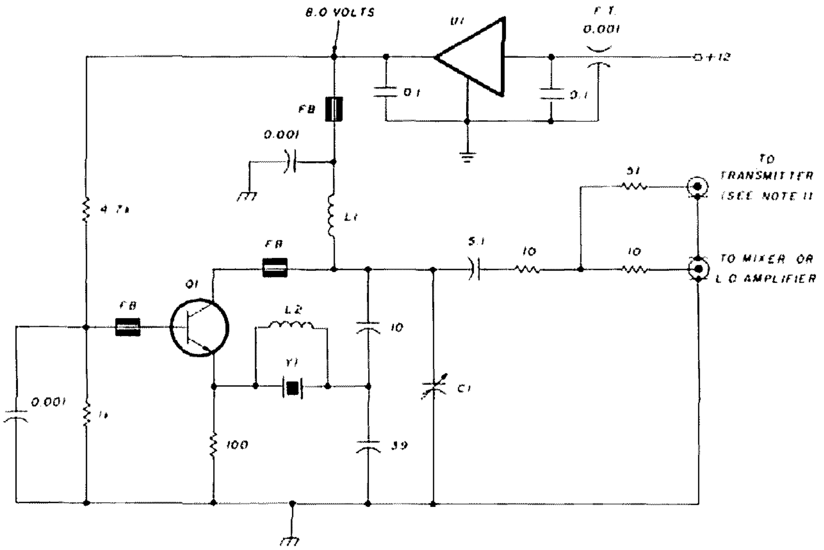
The U310 preamplifier circuit, if used, should be carefully laid out with shielding in mind. First a hole should be drilled near one side of the center of the circuit through the PC board. Its diameter should be 0.191 inches (4.85 mm), the diameter of a No. 11 drill. Next, place the U310 in the hole upside down and quickly solder the gate lead to ground as well as the tab on the U310 can. This provides additional input to output isolation. Next place a shield with a notch for the transistor leads across the JFET to isolate the input and output circuits as shown in fig. 2.

Leads in the oscillator should be kept short, especially those going to the crystal and its associated components. Also, it's wise to place the oscillator in a box separate from the LO amplifier so that any heat generated by the amplifier will not affect the stability of the oscillator. This has a secondary benefit: if the oscillator is separate, it can be used directly with a standard (+5 dBm or 7 milliwatt) DBM until you upgrade to the high-level DBM.

tune-up

First connect all the modules together as a full converter per fig. 1. Then connect an appropriate 2-meter antenna to the input of the converter. If the oscillator is functioning, noise and possibly signals will be heard.

The local oscillator should be tuned for maximum output as indicated on an RF power meter. *Don't detune the oscillator for frequency "netting"; this*



C1	1-10 pF low-loss trimmer capacitor
FB	ferrite bead (Ferroxcube 56 590 65/3B or equivalent)
U1	78L08 8-volt three-terminal voltage regulator
L1	8 turns No. 24 AWG enameled wire close wound on a 0.1 inch (2.5 mm) ID
L2	0.39 μH RF choke or 10 turns of No. 28 AWG enameled wire on a T-25-6 toroid
Q1	2N5179, NE73432B, or equivalent transmitter
Y1	5th overtone series mode 116 MHz crystal in HC18/u wire leaded holder

fig. 7. Recommended oscillator circuit for a 2-meter HDR receive converter. The values shown are for 116 MHz and can be used over a wide range of frequencies. If an oscillator is desired over a larger range (perhaps ±25 MHz) the coil, L1, can be increased or decreased slightly to obtain resonance.

will decrease frequency stability. If only a diode detector is available, place a 3 to 10-dB attenuator pad on the output of the oscillator and the diode detector on the output of the pad. Tune the oscillator for maximum output power.

If a normal range DBM is used, connect the LO directly to the DBM. If an HDR DBM is used, connect the LO to the LO amplifier with a piece of coaxial cable and measure the available output power. It is not critical, but should be between 50 to 150 milliwatts for the HDR mixer configuration.

If the lossless feedback preamplifier is used, it doesn't require tuning but the input and bandpass filters do. The input filter can be peaked on a weak signal or tuned for minimum VSWR on a low-power test set-up. The bandpass filter is best tuned on a sweep set-up, but again could be tuned for maximum signal strength without a great loss in performance.

If the U310 preamplifier is used, first set the input capacitor for nearly minimum capacitance. Next peak the output capacitor for maximum gain or noise when in a typical circuit. The typical gain should be 11 to 12 dB. If noise figure measurement gear is available, tune the input circuit for the best noise figure. Sometimes this can be done at local VHF/UHF conferences, especially when noise figure measurement contests are conducted.

performance

At first, performance may not seem spectacular. In fact, the gain may seem low despite the noise figure available. If you can borrow a noise figure meter, check the performance so you'll know you're where you should be. You should be able to hear just

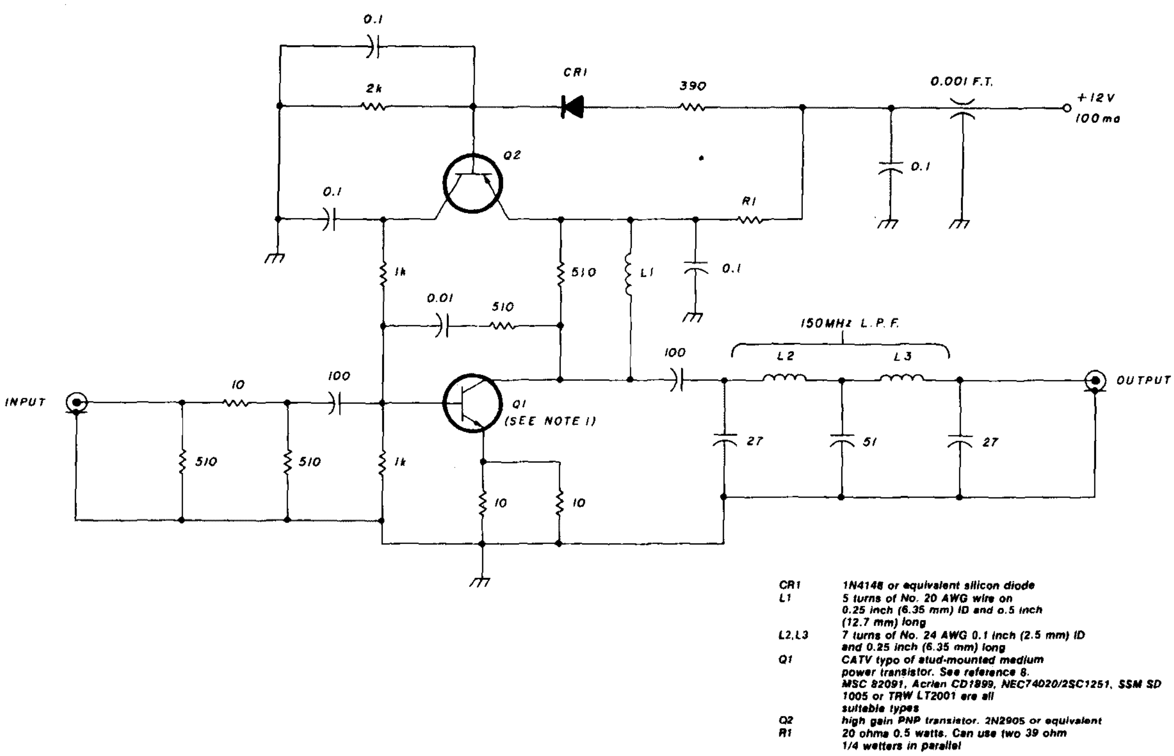


fig. 8. Recommended local oscillator amplifier for use with high-level DBM. ***Note 1:*** **Bolt Q1 stud to cover of box to provide better heatsinking.**

about anyone who's working into your area, if your antenna gain is sufficient.

In the presence of strong signals, you will immediately notice the improved performance. Now you'll see why I've previously stressed that the IF will ultimately be the limiting device! If you used the modular approach, the circuits can be swapped at will as you step up in performance. Also, should you ever suffer a failure, repair or bypassing is easily accomplished. It may even be smart to incorporate a bypass switching arrangement to switch out the IF postamplifier when high level signals are present.

summary

This month's column again stressed improved performance that is within the means and skill level of most Amateurs. Those who have tried these approaches have been pleasantly surprised. It's nice to know that you're keeping up with the state-of-the-art! When good gear is used, contacts are more enjoyable and the weak ones are easier to work.

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3. Joe Reisert, W1JR, "VHF/UHF World: Low-Noise GaAs FET Technology," *ham radio*, December, 1984, page 99.
4. Joe Reisert, W1JR, "VHF/UHF World: Improving Meteor Scatter Communications," *ham radio*, June, 1984, page 82.
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6. M. Martin, DJ7VY, "A New Type of Preamplifier for 145 MHz and 435 MHz Receivers," *VHF Communications*, Spring Issue, 1978.
7. Joe Reisert, W1JR, "VHF/UHF World: VHF/UHF Receivers," *ham radio*, March, 1984, page 42.
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important VHF/UHF events

November 2:	*Peak of Taurids Meteor Shower predicted at 0930 UTC*
November 2-3:	*ARRL International EME Contest*
November 3:	*Peak of Casseopids Meteor Shower predicted at 0930 UTC*
November 12:	*EME Perigee*
November 17:	*Peak of Leonids Meteor Shower predicted at 0300 UTC*
November 23-23:	*ARRL International EME Contest*
December 2:	*7-11 PM Local, 2-meter SWOT Contest (contact K5IS for details)*
December 11:	*EME Perigee*
December 13:	*Peak of Geminids Meteor Shower predicted at 0650 UTC*
December 21:	*Peak of Ursids Meteor Shower predicted at 2200 UTC*
December 21:	*± 1 month, winter peak of sporadic-E propagation*

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

the mystery of the tapered element

I received an interesting letter in the mail the other day. It seemed some of the DXers were having a lively discussion on the effects of taper on the elements of a Yagi beam antenna. Recently published information on tapered elements seemed to indicate that a severely tapered element could actually be longer than a half-wavelength yet still be resonant.[1]

The letter writer concluded that it was reasonable to see that a "thick" element is shorter than a "thin" element, but it was beyond the realm of possibility that a tapered element could be longer than either a "thin" or "thick" element.

My friendly inquirer closed his letter by saying that a "longer-than-normal resonant element of the tapered variety contradicts the laws of Nature. Say it isn't so!"

the background

This was an interesting letter. *Is* it possible to have a tapered element physically longer than a half-wavelength at a given frequency, yet resonant at that frequency? A little insight into basic antenna theory might provide the answer. To quote the *ARRL Antenna Handbook*:[2]

The shortest length of wire that will resonate to a given frequency is one just long enough to permit an electric charge to travel from one end to the other and then back again in the time of one RF cycle. If the speed at which the charge travels is equal to the velocity of light, 300,000,000 meters per second, the distance it will cover in one cycle will be equal to this velocity divided by the frequency in cycles per second, or

$$\lambda = \frac{300{,}000{,}000}{f_{(Hz)}}$$

in which λ is the wavelength in meters. Since the charge traverses the wire twice, the length of wire needed to permit the charge to travel a distance λ in one cycle is λ/2, or one-half wavelength.

Since the speed of light is a universal constant, either the frequency can be adjusted to a given wire length, or the wire length can be adjusted to a given frequency. Finally, by changing from the metric system to the English system and dividing the formula by two, the familiar and useful formula for a half-wavelength in space is:

$$L = \frac{492}{f_{(MHz)}}$$

where L is length in feet of a half-wavelength for a frequency, f, in MHz.

The final step is to take into account the ratio of length to diameter of the conductor. The smaller the ratio, the shorter the antenna will be for a given electrical length. This is illustrated for tubing elements in the graph fig. 1.

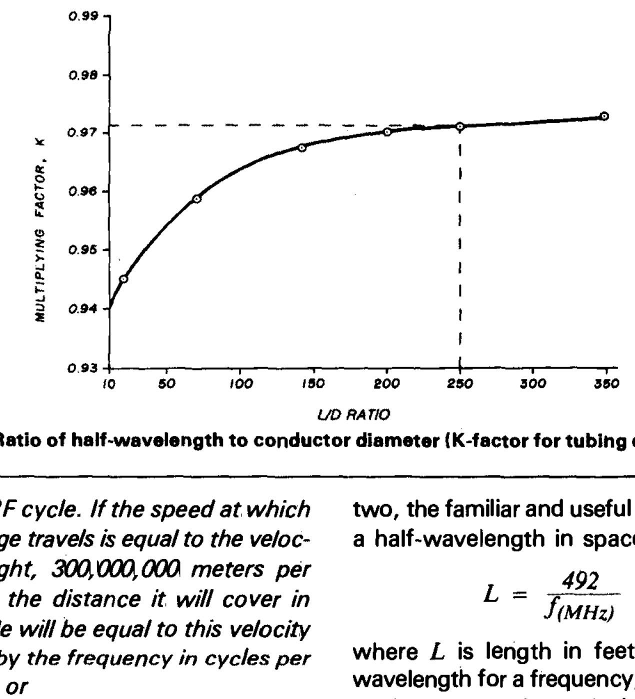

fig. 1. Ratio of half-wavelength to conductor diameter (K-factor for tubing elements).

A length-to-diameter ratio of about 4,000 to 1 is appropriate for a wire antenna in the HF range. A length-to-diameter ratio of, say 250 to 1 could apply to a resonant element made of aluminum tubing. The tubing element would be about 97 percent as long as a wire element for the same frequency. It is possible, then for a "thick" element to be quite a bit shorter than a

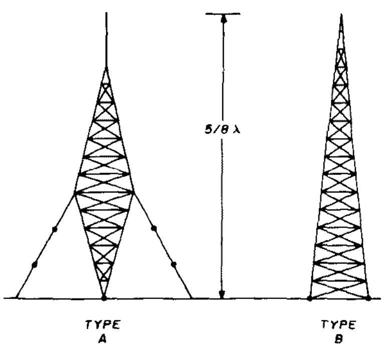

fig. 2. Early broadcast antenna towers.

"thin" wire element, or than the free-space dimension of a radio wave of a given frequency. No argument so far, is there?

what the broadcasters found out in 1934

In 1924 Stuart Ballantine showed that, for a given amount of radiated power, the field strength at the horizon would be greatest when the vertical antenna (over a good ground) was 0.64 wavelength high.[3] This was the concept behind the popular 5/8-wavelength vertical antenna.

This was of immense benefit to broadcast stations because it allowed them to have a signal about 40 percent greater than that provided by a 1/4-wave vertical antenna fed the same power. During the period between 1925 and 1930, many broad-cast stations in America and Europe switched over to the new antenna design.

In attempts to obtain this optimum situation the broadcaster had to erect a tower more than twice as high as had been used previously. Two types of tower designs were available, as shown in **fig. 2**. The type A installation consisted essentially of two towers placed base-to-base and held in a vertical position by four or more guy wires. The type B design was more conventional, with the four tower legs mounted on base insulators. No guy wires were required.

After using these towers for some months it became apparent that the results achieved were not consistent with theory. The promised gain failed to materialize, and the unwanted, high angle radiation was not appreciably reduced. In May, 1934, extensive tests were run on the radiation pattern of WABC (in Wayne, New Jersey), using an airplane to provide vertical pattern plots. The results were discouraging (**fig. 3**).

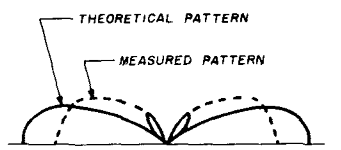

fig. 3. Airplane field strength measurements on WABC antenna showed radiation pattern was not what was expected.

The next step was to measure the current distribution along the tower. In the case of the type A design, the 5/8 wavelength tower should have the current distribution shown by curve *A* of **fig. 4**. But the measured current actually resembled curve *B*! The current distribution did not resemble the sinusoidal curve predicted by theory, nor was there a current reversal approximately a 1/2-wavelength from the top end of the antenna! The top portion of the antenna carried very little current and the bottom portion of the antenna nearest the ground carried most of the current.

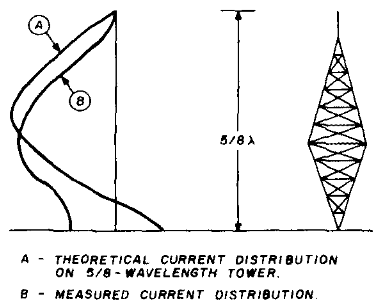

fig. 4. (*A*) Theoretical current distribution on 5/8-wavelength tower and (*B*) measured current distribution.

H.E. Gihring and G.H. Brown of the RCA Victor Company ran tests on antenna models built to scale for a wavelength of 4 meters. The tests on the models duplicated the results found by measurement on the larger broadcast antennas.[4] The clue was that the current distribution on the tower was *non-sinusoidal*, and the rules that applied to normal antennas with sinusoidal current distribution did not apply to towers of irregular cross section!

Armed with this information, tests were run on the type A tapered antenna tower at WCAU (in Philadelphia, Pennsylvania). This tower has an additional 100-foot shaft protruding out of the top. They found little current in the shaft and, again, no reversal of current at the half-wave point along the tower.

The final check was to build a 4-meter model antenna out of wire having a constant cross-section. They reported "substantial agreement" between theoretical and observed values.

problem defined how to solve it

Gihring and Brown had proved that it would be desirable to make the cross-section of the 5/8-wave antenna constant if it were to obey the theoretical laws and deliver the anticipated power gain. They proved this in simple fashion. They made a wooden framework the same size as the maximum tower cross-section for the WCAU tower and dropped wires down to the corners of a square frame placed at the base of the tower (**fig. 5**). Voila! Even with as few as four wires, the current distribution on the tower approximated the desired sinusoidal waveform and the radiation pattern proved to be what was predicted for the 5/8-wave antenna height.

So there it was. By summer, 1935, broadcasters had started to shift away from self-supporting towers in favor of uniform cross-section, guyed towers. And these are the tower designs that are in use today by the majority of broadcast stations around the world.

the taper effect

Although they had a different goal

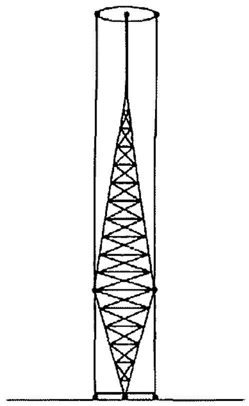

fig. 5. Wire frame suspended from "hat" atop WCAU tower restored proper current distribution along the tower.

in mind, Gihring and Brown had defined and described the so-called "taper effect" — that is, that normal assumptions about antenna length do not apply when the antenna element is not unifrom in cross-section throughout its length. The antenna can be "thick" or "thin" and all is well, but when it's tapered, all sorts of strange things seem to happen!

the discoveries of W3MWC and W6KPC

While the tapered tower effect was well-known in the broadcast industry, it remained unknown to the Amateur fraternity in general. The taper problem did not apply to VHF beam antennas, it seems, because most of them were made of a single section of tubing for each element and no taper was present. Even conventional 6, 10, and 20 meter beams had little, if any, taper. For experimenters who built HF beams with tapered elements, the puzzling results and poor beam performance could not be linked to the conventional element dimensions in use.

As far as I know, the first discussion of element taper in Amateur literature was presented by my good friend Frank Clement, W6KPC.[5] Frank found that his 14 MHz driven element ended up 17 feet 2 inches (5.23 meters) long because of the extreme taper. (Note that the element is longer than the conventional electrical half-wavelength.)

There the matter rested until 1967, when Jim Berger, W3MWC, attempted to build a 3-element, 40-meter beam with tapered elements. In a letter to *QST*,[6] Jim noted that he had to lengthen his elements to make the antenna work properly. The tapered driven element ended up 71 feet (23.6 meters) long (again, much longer than the conventional electrical half-wavelength.)

W2PV defines and solves the problem

Aided by data from W6KPC, I derived a simple chart that provided a correction factor for a tapered element, based upon the maximum and minimum diameters of the element. It proved to be practical, and a few beams with tapered elements were designed from my data with good results. However, I had no mathematical proof that my supposition was correct.

I pushed the matter to the back of my mind until the late Jim Lawson, W2PV, published a mathematical explanation of element taper in his monumental series of articles in *ham radio*.[7] The computer program he suggested was quickly compared against my heuristic (cut-and-try) data and I was pleased to find excellent agreement. A variation of the original program has since been published in *ham radio*.[8,9]

and the answer is

Yes, it is entirely possible for a "half-wave" element to be longer than predicted by conventional formulas *if* the element in question has a nonsinusoidal current distribution along it. One of the most common examples (there are others) is the simple tapered element having a non-uniform diameter along its length. But the means are now at hand to predict this aberration and to compensate for it.

the 80-meter Yagi at OH1RY

Build an 80-meter Yagi antenna?

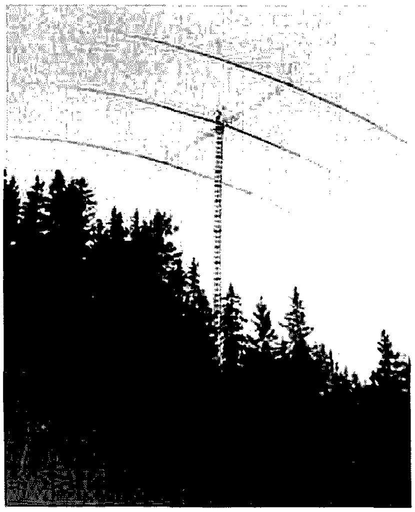

fig. 6. The 80-meter Yagi of Peter, OH1RY. If you look closely, you can see Peter standing atop the center of the beam.

Impossible. But Peter Kolehmainen, OH1RY did it! The antenna is shown in **figs. 6** and **7**. This monster is atop a 92-foot (30.6 meter) high tower placed on the crest of a 60-foot (18 meter) hill, making the beam about 150 feet (46 meters) above the surrounding territiory. The boom length (including the tip guying supports) is 72 feet (22 meters) long. The antenna itself weighs about a half-ton (454 kg).

Element taper? The element is 6 inches (15.24 cm) in diameter at the boom, tapering to 1/2-inch (1.27 cm) diameter at the tips! As you can imagine, this provides some bizarre element lengths. The driven element is 135.5 feet (41 meters) long for resonance at 3.8 MHz. The first photo shows Peter squatting on the driven element, which is hinge-mounted to heavy insulating plates bolted to the side of the boom. The weight of the element is supported by the small "mast" and guy wire assembly behind him (I wonder where the photographer was standing when this picture was taken?)

If you look at the photo of the beam, you'll see Peter standing atop the antenna, dwarfed in size by the monster he has created!

Does the antenna work? Just listen for OH1RY and judge for yourself!

fig. 7. Peter, OH1RY, atop the driven element of his 80-meter Yagi. Center of element is 6 inches (15.24 cm) in diameter!

good reading from Canada

Our neighbor to the north, Yuri Blanarovich, VE3BMV, is the editor of a publication that all Amateurs (especially DXers) should be aware of. A first-class production *Radiosporting* covers all aspects of contest and DX operating — antennas, equipment, station operation, stories of DXpeditions — the works! I enjoy every issue of this publication. If you're interested in it, contact Yuri at Box 65, Don Mills, Ontario, Canada M3C 2R6. (I wish somebody would explain the British and Canadian postal code system!)

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ham radio

PRACTICALLY SPEAKING ...

Joe Carr K4IPV

tracking the hideous intermittent: part 1 — mechanical intermittents

Repair problems are never fun, especially since they take us off the air more times than not.

Probably the nastiest, meanest, and most contemptible of all repair problems are the intermittents. These come and go, usually occurring when it's least convenient. They almost never occur when you're ready to troubleshoot; you stand there, instrument probes in hand and brain engaged, only to find the dam set working properly. You turn your back, set the probes down, and — *zot!* — the trouble reappears momentarily and then disappears again.

Although I'm the first to admit that finding intermittents sometimes falls into the unholy realms of sorcery and witchcraft, certain things can be done to enhance the probability of success.

There's no such thing as a universal procedure for locating intermittents because different kinds of equipment have somewhat different requirements. To keep things simple, we'll limit the scope of this discussion to troubleshooting a high-frequency SSB transceiver with receive problems.

where to look

One thing you can do is educate yourself about the kinds of parts and faults most likely to create intermittent symptoms. Very high on the list are switches and relays. Because these devices are mechanical, they're subject to wear and tear. You'll find dirty internal electrical contacts, poor spring tension, and other faults causing intermittent symptoms. In many cases, a session with a pencil eraser on the contacts or a squirt of contact cleaner (for example, *Blue Stuff*) will work wonders. In other cases, however, only replacement will solve the problem.

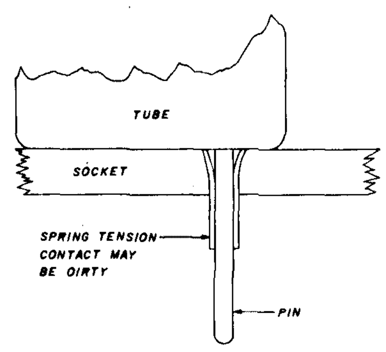

fig. 1. Dirty tube sockets create intermittent conditions.

Another potential sore spot is potentiometers. These components are variable resistors in which a shaft-operated wiper electrode rubs against a wirewound or carbon resistance element. If either the element or electrical contact on the wiper gets dirty, then operation can become intermittent. Unless the dirt has physically damaged the resistance element (as sometimes happens, especially on carbon elements), a simple squirt of contact cleaner will solve the problem. Be especially aware of potentiometers that normally pass direct current (DC) through the wiper connection. I recall a 1963/64 car radio model in which cost-conscious engineers eliminated a coupling capacitor from the volume control and audio preamplifier circuit, thereby making the volume control resistance part of the preamp transistor's bias network. Passing the DC bias through the control generated a massive warranty problem for the manufacturer as those volume controls were chewed up by the truckload!

The printed circuit board (PCB) is another common source of intermittent problems. Two forms of the problem are common: poor solder joints and damage, sometimes hidden, to the board. Both types of fault are especially aggravated in hot parts of the equipment: near power transistors, rectifiers, vacuum tubes, power resistors (2 watts and above), lamps and so forth. These areas sometimes can be identified by discoloration of the PCB. (We'll discuss PCB problems in a moment.)

In vacuum tube equipment, the tube socket can produce intermittent problems. If the tube pins or the socket contact (fig. 1) lose tension, then an intermittent connection results. These faults can be repaired in most cases. If dirt is the problem, remove the tube and gently clean its pins with a dime store ink eraser, spray the pins with a clear contact cleaner, and reinsert the tube into the socket. Next, pull the tube out of the socket and then reinsert it four or five times in a row. This action will clean the socket. Wait a hal

hour or so for the cleaner to dry, then turn the rig on to evaluate the results. If the intermittent remains, re-tension the socket contacts with either a tiny screwdriver tip or other sharp-pointed tool.

Components can be the source of some maddening intermittents. Unfortunately, some component problems tend to heal themselves the instant a probe is attached (especially semiconductors). Be especially wary of plastic packaged transistors, tubular (non-mylar) capacitors, and resistors. These components account for a large portion of the problems.

Shielded IF and RF transformers (**fig. 2A**) are frequent sources of intermittent faults. The coil wire attaches to the lugs on the base, and these sometimes break (see **fig. 2B**). In some cases, a careful worker can repair these transformers; it's merely a matter of resoldering — you'd be surprised how many escape the factory unsoldered!

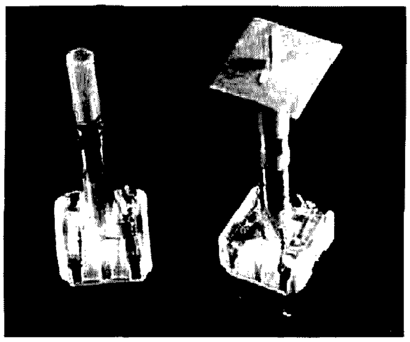

fig. 2A. RF/IF transformers.

fig. 2B. Another cause of intermittency is traceable to broken coil lead/base LUG connections.

finding the intermittent

The first step, crucial to quick success, is observation. Define in your mind what the rig is doing wrong, what functions are affected, and whether it happens on both receive and transmit. Narrowing down the possibilities allows you to restrict your efforts to a certain few stages, once you determine which stages may or may not be affected. For example, if the problem happens on SSB but not CW, or on receive but not on transmit, we can then infer that the problem is probably not in a stage that is common to both affected modes. Deciding which stages are likely candidates depends on understanding your transceiver; do a block diagram analysis and read any circuit descriptions provided by the manufacturer.

Some intermittents occur under vibration, touching, or thumping. A certain number of such problems are due to bad switches and potentiometers. A little *light* tapping with an insulated probe, a little jiggling, or visual inspection will often locate the source of the problem.

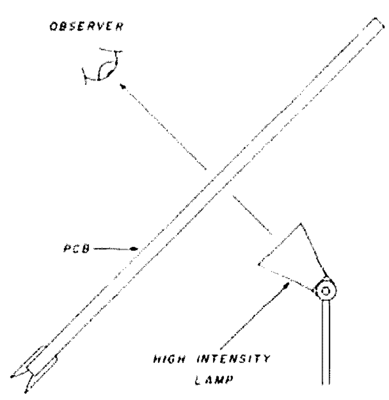

fig. 3. View the PCB with light shining through to find cracks.

Most mechanical faults are due to either bad PCB joints or bad components mounted on the PCB. Unfortunately, many of these problms fail to yield to the tapping method because any vibration at all, anywhere on the board, will produce the fault. Two approaches helpful in this regard don't require formal troubleshooting: visual inspection and shotgun solder touch-up.

Visual inspection involves examining every joint on the PCB with a 10X or so magnifying glass under adequate light. Examine the board two ways: first with the light shining on the soldered side, then with the light shining *through* the PCB from the component side (**fig. 3**). In the latter case, subsurface cracks in the PCB material can break a joint or track. Even if the joint or track appears normal it should be reworked.

Visual examination takes a certain amount of practice; one needs to develop a "small eye," that is, the ability to see defects where others would see a "normal" joint.

I usually inspect PCBs with a bottle of fingernail polish or a grease pencil handy. Especially on large boards, each apparent anomaly is marked so that I can find it easily later on. This habit is especially useful when using a magnifier because the glass will distort your perception of space.

Shotgun soldering is especially useful when the area of the intermittent is known, when the PCB is small, or when nothing else seems to work. I can recall another mobile radio receiver problem in which the VHF front-ends PCB had a high "bad joint" intermittent rate, but were difficult to remove and replace. In that case, the more elegant "visual inspection" method was not cost effective, so we pulled the PCBs, soldered every joint, and tinned every track. Rarely did this method fail on that particular problem.

At this point let me digress a little bit to answer the purists who would criticize this approach. I admit that the elegant method is to find the single bad joint or broken track and repair only that. Unfortunately, this approach can be time-consuming and may even be impossible. While the purist "supertech" is messing around trying to analyze which joint is bad, I'm going to fix the rig! Commercially, the shotgun approach is more profitable — and to Amateurs it means getting back on the air sooner.

Next month: Tracking down thermal intermittents.

ham radio

digital frequency readout

using the Commodore 64

A ÷ by 16 prescaler designed for the Argosy I can be adapted to other transceivers

Although digital frequency readout is clearly a major advance in Amateur transceiver technology, many excellent rigs that offer most of the other advantages typical of state-of-the-art gear don't have digital readout. Yet it seems ridiculous to replace a good piece of equipment because of this one shortcoming. If you have such a rig, you might want to consider adding a digital frequency display as an accessory.

This can be done through the purchase or construction of a frequency counter, but such gadgets tie up both station space and dollars, and without special hookups can display only transmitted frequencies. Furthermore, when the station already contains a computer and monitor, it seems redundant to add yet another display with its associated batch of electronics.

I was doing a good job of break-in CW with a Ten-Tec Argosy and a Commodore 64 serving as a CW keyboard. With conditions becoming more and more difficult as the sunspot number fell, the need for sharp filters and "on-the-nose" schedules became too acute to be ignored, and the analog dial of the Argosy fell short of what was needed. Accordingly, I decided to make the computer display the required information.

Fortunately, the C64 contains an array of hardware timers that operate independently off the central processor and can be accessed through the user port. It also contains a stable frequency reference in the form of the machine clock. The timers perform the comparison of unknown and reference frequencies. The mathematics necessary to display the operating frequency, formulated in BASIC, run in real time in order to provide a continuous readout. The system is easy to implement and debug, performs beautifully, and is readily adaptable to other popular transceivers.

timer

The timers are part of two 6526 Complex Interface Adapter (CIA) integrated circuits contained in the machine. One of the two ICs is dedicated primarily to the keyboard interface, but its timers are partly accessible through the user port. The second CIA is dedicated to the user port and the serial interface, and its functions are fully accessible.

The timers are 16-stage binary counters that can be preset to any count between 1 and 65,535. They count down from the preset value and deliver an output when the count passes zero. Thus, they can divide an input frequency by any integral number within their range. Furthermore, they can be cascaded by appropriate instructions to form a longer counter chain.

The signal counted can be either an external input or the machine clock. The latter runs at 1.022727 MHz and is derived from a master crystal oscillator that also generates the color subcarrier in a TV.

The counting rate is limited to about 500 kHz by the internal mechanics of the interface chip. It is therefore necessary to bring the frequency to be measured down into the range below 500 kHz. However, an aliasing effect permits moderate extension of the limit with appropriate modification of the programmed mathematics.

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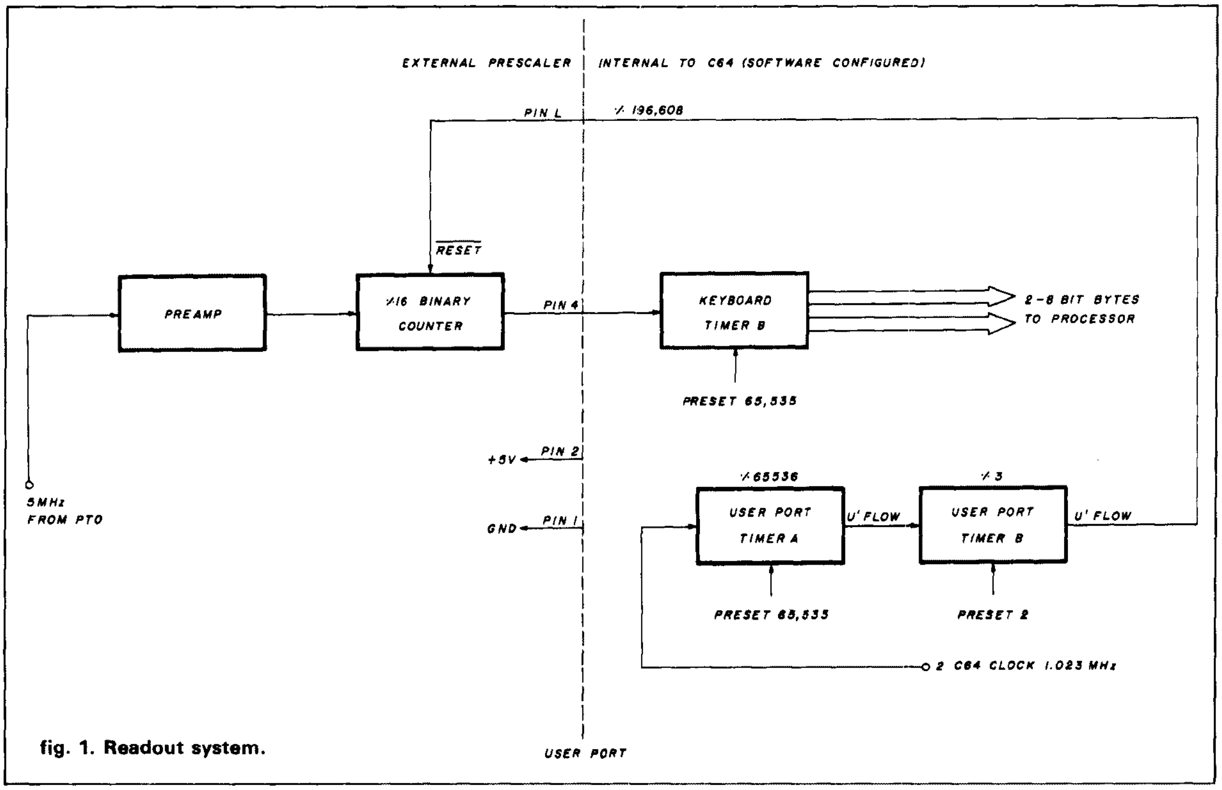


fig. 1. Readout system.

design considerations

A key requirement for digital frequency readout is precise gating of the counter that records the frequency to be measured. This would be difficult to do if the gating were done through software, since the machine cycle times for executing BASIC instructions are a significant fraction of the counting period required. Fortunately, the designers of the CIA arranged for direct access of the timer output through the user port, without intervention by the computer's processor; thus, one timer can be used to precisely gate another by the addition of simple external circuitry, and the result can be examined by the computer at its leisure.

In general, neither the transmitted nor the received frequencies are present as a steady-state signal (except in the case of a direct-conversion system), so an indirect method of determining the frequency must be used. Since most modern gear uses a master oscillator covering a fixed range, in conjunction with crystal-controlled heterodyning oscillators for the various bands, it is generally satisfactory to measure the master oscillator frequency and to apply appropriate corrections to the display. The stability of the heterodyning system is usually good enough to make the correction a constant factor, and no significant error results from warm-up or temperature changes.

My Argosy oscillator operates in the 5-MHz region, as do the oscillators in many other transceivers. To get the frequency below the 500-kHz limit, the frequency to be measured must be reduced by a factor of better than 10. Translation through a mixer could be used, giving an output of 0 to 500 kHz as the oscillator covers 5 to 5.5 MHz, but an additional crystal oscillator would be needed and the mixer output circuit would have to be broadbanded over the range from DC to 500 kHz.

It's easier to use a frequency divider in the form of a simple integrated-circuit counter. For example, a four-stage binary counter divides by 16, giving an output frequency range of 312.5 kHz to 343.75 kHz as the input varies from 5 MHz to 5.5 MHz.

An obvious disadvantage of the divider is the loss of precision associated with the compressed frequency range. The precision can be regained, however, by lengthening the counting interval by an equivalent factor. Normally, achieving 100-Hz resolution — which is typical of most digital readout transceivers — would require counting for 0.01 second; with the divide-by-16 prescaler, the time must be lengthened to 0.16 second or more. Fortunately, a repetition rate of two or three

times per second is fast enough for a useful real-time readout, so the divider scheme is practical. It is easy to implement, requiring only the binary counter and a transistor buffer-driver to amplify the small amount of RF taken from the 5-MHz transceiver oscillator. Power can be taken from the computer's user port.

The computer program for performing the readout and display handles a number of functions. First, it arranges the timers for the task at hand. Then it starts the count cycle running and monitors for completion of the cycle. Next it obtains the resultant count and performs the mathematics necessary to deduce the oscillator frequency from the count. It applies the appropriate corrections for the band and mode in use, as input by the operator. Finally, it displays the resultant figure on the screen, and initiates a new cycle. Despite all this activity, the program is a short one, requiring fewer than 50 lines of typing and only a few seconds to load from tape or disk.

system implementation

The readout system is shown in **fig. 1**. Timer *A* of the user port divides the machine clock frequency by 65,536. When its count passes zero a pulse is delivered to user port timer *B*. The latter is preset to a count of 2; when it also indicates an underflow (at a clock count of $65,536 \times 3 = 196,608$), line PB7 (Pin L) of the user port switches from plus 5 volts to zero volts. This voltage transition gates off the external counter. The counting interval is $196,608/1,022,727 = 0.192239$ second.

Counter *B* of the keyboard interface CIA counts the input pulses delivered from the external divide-by-16 prescaler through Pin 4 of the user port. It is preset to 65,535 at the start of each cycle and counts down until the external signal is gated off.

The schematic of the external prescaler is shown in **fig. 2**. The binary counter is a TTL 74197, which is capable of 70-MHz operation. There are numerous other possible choices, including some CMOS varieties; however, with the latter it may be necessary to run the prescaler at a higher voltage than 5 volts to achieve the 5 MHz counting speed. This would in turn lead to power supply and interfacing complications.

A cursory glance at the transistor buffer circuit suggests that it violates all the rules of stable biasing for amplifiers. So it does, but the objective is not linear reproduction of the signal, but rather the development of a square wave drive from a hundred millivolts or so of sinusoidal signal. Thus, it is permissible to operate the stages at a small degree of either saturation or cutoff. The resistor values shown keep the 2N2222s in such a state.

The coupling to the transceiver is through a simple capacitive pickoff and a 3-foot (0.9 meter) shielded cable to the prescaler. The loss in signal caused by the voltage-divider effect of the cable capacitance is made up by the two-stage amplifier in the prescaler.

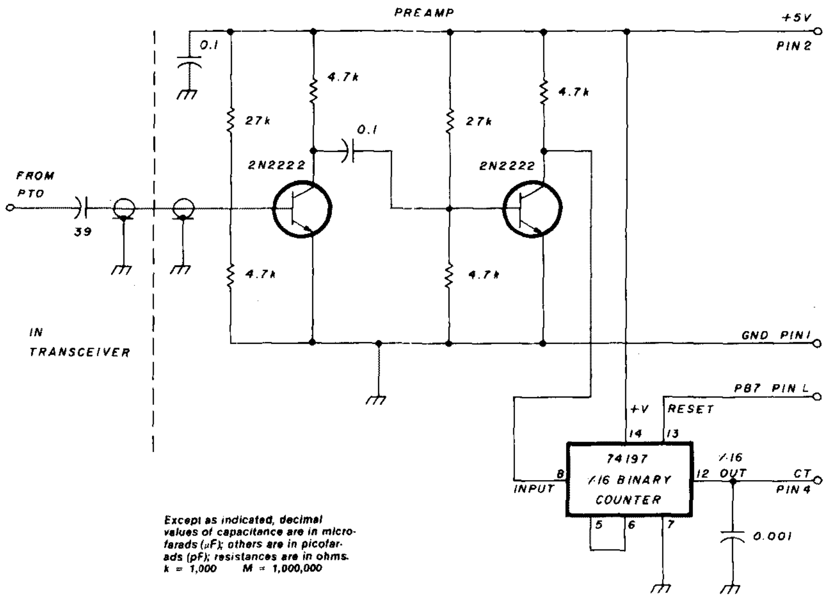


fig. 2. Prescaler schematic. Bias resistor values are chosen for a small amount of saturation or cutoff. Stage delivers square wave drive from about a hundred millivolts of sinusoidal signal.

The advantage of this approach is that it requires no active hardware inside the transceiver and produces no large-amplitude square waves near the transceiver's receiver circuitry.

In my Argosy, the 39 pF capacitor is connected to the output (rearmost) terminal of the PTO assembly and a shielded cable is run from there to one of the spare RCA jacks on the rear panel. Many transceivers have the VFO signal already available on the back panel and require no internal modification. In general, the pickoff point should be at the output of the VFO buffer, and the rig should be checked after connection of the prescaler to make sure that there is no loss of output.

The prescaler is located close to the user port and enclosed in a small aluminum box (fig. 3) to prevent radiation of harmonics of the scaled-down frequency every few hundred kilohertz throughout the spectrum. Such radiation is also suppressed by the 0.001-μF capacitor from the prescaler output to ground, which slows down the edges of the square wave going into the port.

At W3NNL, the prescaler shares the user port with an MFJ-1228 CW/RTTY interface, and is piggybacked on the latter (fig. 4). There is only one port line (Pin L) in conflict, and a switch is provided to transfer it from frequency readout to the MFJ. I have also combined a CW keyboard program (not shown) with the frequency readout.

In the absence of such a piggyback scheme, it will be necessary to procure a connector with the required 0.156-inch (4-mm) contact spacing. These seem to turn up at hamfests only in much longer sizes than the 24-pin variety needed for the user port; however, the excess length is easily removed with a hacksaw.

computer program

The BASIC program shown in fig. 5 can be followed using the memory map presented in the C64 Programmer's Reference Guide. **Lines 110** through **290** are concerned with acquiring the operator-specified band and mode, and defining the pre-established correction factors. The section from **line 300** through **420** sets up the timers, runs the count cycle, and calculates the frequency. The remainder of the program is devoted to formatting and displaying the output and starting a new cycle.

The timer interface uses POKE and PEEK statements directed to the appropriate addresses. The count in timer *B* of the keyboard CIA is picked off in two eight-bit bytes, and the number of pulses is calculated and multiplied by a factor which yields the original input frequency to the prescaler. The low-edge oscillator frequency is subtracted from the input frequency, and the lower edge of the band in use is added. Finally, corrections unique to the individual transceiver are

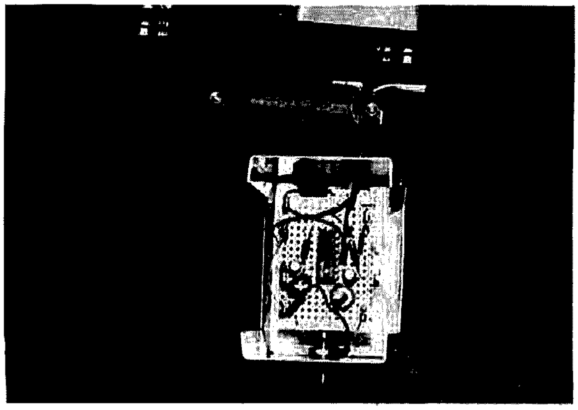


fig. 3. Prescaler is enclosed in a small aluminum box to prevent harmonic radiation.

either added or subtracted; these corrections take care of crystal-tolerance errors in the heterodyne frequencies, deliberately introduced offsets such as those used in the Argosy to avoid birdies, and shifts in the received and transmitted frequencies dependent on the mode of operation selected (CW or either sideband).

The correction consists of a term *K* representing the conversion-oscillator related effects, a factor *D* representing the offset between CW and normal sideband operation, and a multiple of *D* which is the shift incurred when the reverse sideband is selected. Of course, if only one mode is to be used, *D* may be omitted from the selection statements and from the frequency equation of **line 400**, and the mode select routines of **lines 130-160** and **270-280** can be omitted. The value of *K* can be determined for the mode of interest.

Several tasks are associated with formatting and display. The screen is cleared and the "FREQ = " leader is printed, followed by the frequency, a wide space, and the mode entered. To avoid a string of meaningless extra digits after the decimal point, the computed frequency is rounded off to the nearest 0.1 kHz. To eliminate the visually unpleasant sensation produced by the C64's dutiful suppression of the decimal point and following zero on integral kilohertz readings, some jockeying is performed to tack the point and zero back on. Finally, compensation is made for the leading-zero suppression, which occurs when tuning from above 10 MHz, to below 10 MHz, and which would otherwise shift the display left and expose a spurious right-hand digit.

The above action is repeated every few hundred milliseconds, except that the screen is left uncleared after the first cycle. Clearing produces flicker and is unnecessary unless the format changes.

For the display to work properly, it is imperative that

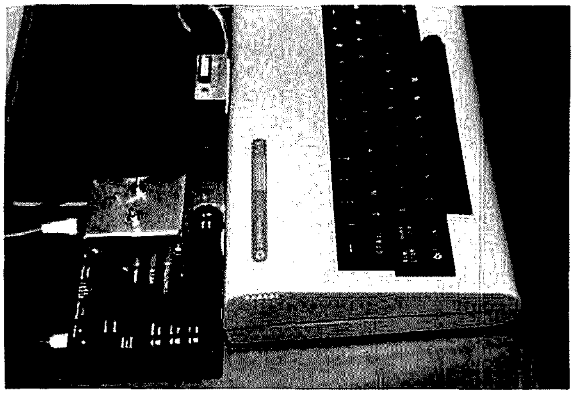


fig. 4. Prescaler mounted on MFJ interface.

the display portion of the program be typed *exactly* as specified, including all embedded punctuation marks.

correction factor determinations

When the program is first typed, **lines 180** through **280** should be omitted. The missing lines may be entered after determining the *K* and *D* factors as described below.

If the correct nominal low-edge oscillator frequency has been entered on **line 290**, the indicated frequency will be somewhere within 10 kilohertz or so of the true value when the program is run and WWV, CHU, or the crystal calibrator is tuned in. Before starting the correction factor procedure, the calibrator should be checked to make sure it is accurately tuned to the standard-frequency station.

To determine *K* for a given band, only one measurement is necessary. With the rig set in the normal sideband position for the band in use, the crystal calibrator should be tuned to zero beat. The frequency indicated by the display should be noted and subtracted from the actual frequency of the calibrator harmonic. The result is the value of *K*. This operation should be repeated for each band or bandswitch segment and the numbers recorded for entry into the program (don't forget the minus signs where applicable).

Next, the transceiver should be set for CW and the calibrator harmonic tuned for the proper CW pitch rather than for zero beat. The direction and magnitude of the change in indicated frequency should be noted. The change will normally be about 0.8 kHz. An accurate value can readily be determined if the rig has a sharp audio or crystal filter. If a separate receiver is available, a still more accurate measurement can be made by noting the display reading when the transceiver is placed in the transmit mode and zeroed in the external receiver with a known reference. The RIT or OFFSET control should be set to zero for this measurement.

The value of *D* obtained is positive if the indicated frequency decreases as compared to the zero beat setting, and negative if it increases. By way of example, in the program (**fig. 5**) at 10 MHz the indicated frequency with WWV at zero beat in the SSBN (normal sideband) mode was 10010.8 kHz, leading to a *K* of minus 10.8. With the WWV carrier peaked in the audio filter, the indicated frequency increased another 800 Hz to 10011.6 kHz, yielding −0.8 for *D*. The numerical value of *D* holds for all bands, but the sign changes as the rig implements CW on different sides of zero beat.

Finally, the opposite sideband should be selected and the zero-beat procedure followed. The correction determined should be divided by D to give the multiplier used in **line 280**. Again with reference to **fig. 5**, my display read 10014.2 in the reverse (USB) mode, or 3.4 kHz higher than SSBN. This is 4.25 times as large as the CW correction and in the same direction, so the multiplier in **line 280** becomes 4.25.

operation

The program asks for entries of the band in MHz and the mode (0 for CW, 1 for normal SSB, and 2 for reverse SSB). It then displays the frequency to the nearest 0.1 kHz and the mode in use. A flickering of the display between two adjacent tenths digits indicates that the frequency is approximately halfway between the two values.

The readout follows the incremental tuning on receive and shows the transmitted frequency when the key is pressed or the mike activated. If there is considerable RF in the station the readout may be erratic on transmit, but will work if the drive is decreased to reduce the spurious RF.

To change the band and/or mode in use, press the f1 key. This restores the prompts and starts the process again.

adaptation to other transceivers

As described, the program and hardware should work with the Ten-Tec Omni as well as the Argosy. However, a number of other transceivers use upper/lower sideband selection rather than normal/reverse for the band in use. In general, its merely necessary to change the format, substituting LSB and USB for SSBN and SSBR. Proper choice of correction factor values and algebraic signs will then produce equivalent performance.

Of more concern is the VFO frequency and its relationship to output frequency. Both the Kenwood TS520S and the Yaesu FT101E use backward-tuning VFOs' that is, the low-band edge corresponds to maximum VFO frequency. This requires modification of the frequency equation of **line 420** to read:

```
110 REM DIGITAL FREQUENCY READOUT BY W3NNL 3/85
120 PRINT CHR$(147)
130 INPUT "BAND MHZ";B:INPUT"MODE (0=CW, 1=SSBN, 2=SSBR)";M
140 IF M=0 THEN M$="CW"
150 IF M=1 THEN M$="SSBN"
160 IF M=2 THEN M$="SSBR"
170 REM K AND D FACTORS BELOW APPLY TO W3NNL ARGOSY ONLY
180 IF B=3.5 THEN K=-.1:D=-.8
190 IF B=7 THEN K=-.4:D=-.8
200 IF B=10 THEN K=-10.8:D=-.8
210 IF B=14 THEN K=-.4:D=.8
220 IF B=21 THEN K=-10.4:D=.8
230 IF B=28 THEN K=-11.3 D=.8
240 IF B=28.5 THEN K=-11.8 D=.8
250 IF B=29 THEN K=-11.6 D=.8
260 IF B=29.5 THEN K=-11.5:D=.8
270 IF M=1 THEN D=0
280 IF M=2 THEN D=4.25*D
290 E=5000:REM ARGOSY PTO FREQ KHZ AT LOW BANDEDGE
300 PRINT CHR$(147)
310 POKE56589,0
320 POKE56590,1
330 POKE56582,2:POKE56583,0:POKE56591,16
340 POKE56591,71
350 POKE56327,255:POKE56326,255
360 IF PEEK(56577)>127THEN 360
370 H=PEEK(56327):L=PEEK(56326)
380 POKE56335,16
390 POKE56335,33
400 N=256*(255-H)+255-L
410 IF N<32768 THEN N=N+65536
420 F=.08322975*N-E+1000*B+K+D
430 PRINT CHR$(19):PRINT"FREQ =";
440 F=10*F
450 IF F-INT(F)<.5 THEN 470
460 F=F+1
470 F=INT(F)/10
480 IF F<10000 AND F>9000 THEN 560
490 IF F-INT(F)<>0THEN 520
500 PRINTF;:PRINT CHR$(157);
510 PRINT".0";:PRINT M$:GOTO 530
520 PRINTF;:PRINT M$
530 GET C$:IF C$=CHR$(133) THEN 120
540 IF PEEK(56577)<128 THEN540
550 GOTO360
560 PRINT CHR$(29);
570 GOTO 490

READY.
```

fig. 5. C-64 program for digital frequency readout.

420 F = E − .08322975*N + 1000*B + K + D

The TS520 VFO tunes from 5.5 to 4.9 MHz, so the value of E in **line 290** becomes 5500. The FT101 VFO tunes from 9.2 to 8.7 MHz, so E for this rig is 9200.

Since 9200/16 = 575, the FT101 pushes the computer past its upper frequency limit. This is where the previously mentioned aliasing effect comes to the rescue: the increasing rate at which pulses are missed causes the timer to slow down in exact proportion to the frequency excess. It turns out that correct results are obtained if the count is subtracted from 2 × 65536 = 131,072. Thus, **line 410** should be changed to read: 410 N = 131072-N

Rigs that have provision for 160 meters require the insertion of a correction-factor line for that band in the block between **lines 170** and **270**. For the TS520, *B* should equal 1.8, and for the FT101, *B* should equal 1.5. **Line 200** may of course be omitted for rigs which do not have a 10-MHz band.

Both the TS520 and FT101 have external access to the VFO. The TS520 VFO buffer is available through a 2:1 resistive divider at the VFO phone plug connector, and through direct connection at Pin 1 of the EXT VFO socket. The former connection is preferable if enough signal is obtained to drive the prescaler. The FT101 VFO is available at Pins 6 or 7 of the EXT VFO socket.

The program in **fig. 5** will accommodate oscillator frequencies from about 2.7 to 8 MHz. Frequencies below 2.7 MHz can be handled if **line 410** is omitted.

final comments

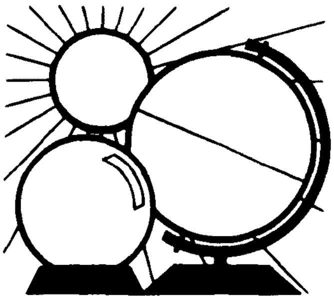
The most likely causes for failure of the system to operate are typing errors in the program or a problem in the prescaler. If the latter is running, it should be possible to hear harmonics at multiples of the divided-down frequency if the receiver antenna is brought near the circuit board. An analog DC voltmeter will read about 2.5 volts when connected to Pin 12 of the 74197. Some experimentation with preamplifier base resistor and input coupling capacitor values may be necessary if the VFO output is very low.

One very quickly becomes addicted to the circuit, to the extent that the rig is never on unless the computer is first fired up. It is satisfying to be able to set the display to 14000 kHz on a seemingly dead 20-meter band and have one of the beacon stations pop out right in the center of the audio passband.

acknowledgements

Thanks are due to Earle Lewis, W3JKX, Fritz Hauff, W3NZ, and Dick Briner, WB3GVU, for furnishing the Kenwood and Yaesu manuals. The photographs were taken by my son-in-law, Herb Hoppe, Jr.

ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

winter DX season

November through February constitutes the winter DX season. Because the D and E regions of the ionosphere receive less energy from the sun in the northern hemisphere during this time, less ionization occurs. Therefore, the daytime attenuation of radio signals is lower in winter than during the rest of the year.

Attenuation is a result of signal energy being absorbed by ions in the D region (35-50 miles or 60-80 km) above the earth. The amount of absorption is related to the zenith angle* to the sun from the points where your path crosses this D region. And on any propagation path, absorption increases with the number of transits of the D region and varies inversely with frequency. So in working DX, it pays to use the higher frequency bands to obtain more distance per hop (resulting in fewer transits) and less signal loss.

This is why we generally think of 6, 10, or 15 meters for DXing. But in winter, particularly near sunspot minium, we have the opportunity to work DX on the lower frequency bands with lower signal loss, day or night, than at any other time of the year. But you can't *always* count on it; signals traveling a high latitude path may be poor for several days at a time. This is known as the winter anomaly.[1]

Along with the lower signal attenuation, the QRN decreases as fewer local thunderstorms pass through your area and the large thunderstorm areas near the equator move further south, requiring more than one hop to get to us. This decreases the noise some 6-8 dB, which is particularly noticeable on the 160, 80, and 40-meter bands.

Even though ion production in the D, E, and lower F regions is lower, ions are better able to diffuse and drift upward along the geomagnetic field lines into the F region. This layer is the major factor in defining the maximum usable frequency. In winter this maximum usable frequency rises rapidly as the sun rises each day, peaking just after noontime, then diminishing during the afternoon, evening, and through the night to a low value just before dawn the next day. The exception to this situation is for locations nearer to the equator, where the ionization continues to drift and diffuse up during the afternoon and evening to become the transequatorial maximums described in my October, 1983 column.[2] The maximum usable frequency peak reached each day and the depth of the predawn minimum frequency of the next morning are related to the solar flux of the day. The higher the flux that day, the higher the frequency peak and the lower the dip the next morning.

Another advantage during the winter season is that the geomagnetic field is least disturbed during November and December. This manifests as least variation of the magnitude and direction of the geomagnetic field lines in an hour's time. This translates into fewer periods of QSB during these months.

last-minute forecast

The first and second weeks of November are expected to favor the higher HF bands, 10 through 30 meters. The solar flux is expected to be higher at this time of the month and result in higher MUFs. If the geomagnetic field is also disturbed at this time then transequatorial propagation on southern paths should also be expected. More hours of darkness, less QRN, and stable signal conditions give an edge to the lower bands for east-west and northern DX contacts this time of year. The lower HF bands are expected to be best the last two weeks of the month. You can update this forecast daily by listening to the time and frequency radio station, WWV, on 2.5, 5, 10, 15, and 20 MHz at 18 minutes after each hour. When the solar flux, as announced, is below 75 and the geomagnetic A is less than 15 or K is less than 4 the lower bands should be best. If the geophysical indices are higher, consider using the higher HF bands instead.

The Taurids meteor showers will occur from October 26 to November 22, with a maximum count of ten per hour from the 3rd through the 10th of November. Lunar perigee is on the 12th, and a full moon falls on the 27th. A total (totality of 1 minute 59 seconds duration) eclipse of the sun is calculated to be visible on November 12th way down in the Antarctic regions. It starts south of Africa at 1209 UT and moves to the tip of South America, ending at 1612 UT. The bands open to Antarctica on the accompanying propagation chart should lower to 40 meters and then recover during the above time period on eclipse day. Try for a contact!

band-by-band summary

Ten, twelve, and fifteen meters, the day-only DX bands, will be open from morning to early evening almost every day, and to most areas of the world. The openings on the higher of these

*The Zenith angle is the angle measured from directly above an observer to the sun (0 degrees when the sun is directly overhead, 90 degrees when the sun is at the horizon).

NOVEMBER

WESTERN USA

GMT	PST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	4:00	30	40	20	10	10	10	10	20
0100	5:00	20	40	20	12	12*	10	10	20
0200	6:00	30	40	20	20*	12	12	12	20
0300	7:00	30	40	20	20	15	15	15	20
0400	8:00	30	40	20	20	15	20	20	30
0500	9:00	40	*40*	20	20	20	20	20	30
0600	10:00	40	*40*	20	20	20	20	20	30
0700	11:00	40	*40*	*30*	20	*20*	*20*	20	40
0800	12:00	40	*40*	*30*	*20*	*20*	*20*	*20*	40
0900	1:00	*40*	*40*	*30*	*30*	*20*	*20*	*20*	*40*
1000	2:00	*40*	*40*	*30*	*30*	*20*	*30*	*30*	*40*
1100	3:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	*40*
1200	4:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	*40*
1300	5:00	*40*	30	15	15	*30*	*30*	*30*	*40*
1400	6:00	*40*	20	12	12	20	*30*	*30*	*40*
1500	7:00	40	20	10	10	15	*20*	*20*	*40*
1600	8:00	40	20	10	10	12	20	*20*	*40*
1700	9:00	*40*	20	10	10	12	20	20	*40*
1800	10:00	*40*	30	10	10	12	15	20	*40*
1900	11:00	*80*	40	12	10	12	12	20	20
2000	12:00	*80*	40	15	10	12	12	15	20
2100	1:00	*80*	40	20	10	12	10	12	20
2200	2:00	*80*	40	20	10	12*	10	12	20
2300	3:00	*80*	80	20	10	10	10	10	20

MID USA

GMT	MST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	5:00	30	40	20	12	12	10	10	20
0100	6:00	30	40	20	15	12	12	12	20
0200	7:00	30	40	20	20	15	15	15	20
0300	8:00	40	40	20	20	20	20	20	20
0400	9:00	40	*40*	20	20	20	20	20	30
0500	10:00	40	*40*	20	20	20	20	20	30
0600	11:00	40	*40*	*30*	20	20	20	20	40
0700	12:00	40	*40*	*30*	*20*	*20*	20	20	40
0800	1:00	*40*	*40*	*30*	*30*	*20*	*30*	*20*	40
0900	2:00	*40*	*40*	*30*	*30*	*20*	*30*	*30*	*40*
1000	3:00	*40*	*40*	*30*	*30*	*20*	*30*	*30*	*40*
1100	4:00	*40*	*40*	15	*30*	*30*	*30*	*30*	*40*
1200	5:00	*40*	30	12	15	*30*	*30*	*30*	*40*
1300	6:00	30	20	12	12	*30*	*30*	*30*	*40*
1400	7:00	40	20	10	12	20	*20*	*30*	*40*
1500	8:00	40	20	10	10	15	*20*	*20*	*40*
1600	9:00	40	20	10	10	12	20	*20*	*40*
1700	10:00	40	20	10	10	12	20	20	*80*
1800	11:00	*40*	30	10	10	12	15	20	*80*
1900	12:00	*80*	40	12	10	12	12	20	20
2000	1:00	*80*	40	15	10	12	12	15	20
2100	2:00	*80*	40	20	10	12	10	12	20
2200	3:00	*80*	80	20	10	12	10	12	20
2300	4:00	*80*	80	20	10	12	10	10	20

EASTERN USA

GMT	CST	EST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	6:00	7:00	40	40	20	15	15	15	15	30
0100	7:00	8:00	40	40	20	20	20*	20	20	30
0200	8:00	9:00	40	40	20	20	20	20	20	40
0300	9:00	10:00	40	*40*	20	20	20	20	20	40
0400	10:00	11:00	40	*40*	20	20	20	20	20	40
0500	11:00	12:00	40	*40*	*30*	20	20	20	20	40
0600	12:00	1:00	40	*40*	*30*	*20*	*20*	20	20	40
0700	1:00	2:00	40	*40*	*30*	*20*	*20*	*30*	*30*	40
0800	2:00	3:00	*40*	*40*	*30*	*20*	*20*	*30*	*30*	*40*
0900	3:00	4:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	*40*
1000	4:00	5:00	*40*	30	12	*30*	*30*	*30*	*30*	*40*
1100	5:00	6:00	30	20	12	15	*30*	*30*	*30*	*40*
1200	6:00	7:00	30	20	10	15	*20*	*20*	*20*	*40*
1300	7:00	8:00	40	20	10	12	20	*20*	*20*	*40*
1400	8:00	9:00	40	20	10	12	15	*20*	*20*	*40*
1500	9:00	10:00	40	20	10	10	12	_30*_	*20*	_80*_
1600	10:00	11:00	40	20	10	10	12	*20*	*20*	_80*_
1700	11:00	12:00	40	20	10	10	12	20	*20*	*80*
1800	12:00	1:00	40	20	10	10	12	15	20	*80*
1900	1:00	2:00	*80*	30	12	10	12	12	20	30
2000	2:00	3:00	*80*	40	15	10	12	12	15	20
2100	3:00	4:00	*80*	40	20	10	12	10	12	20
2200	4:00	5:00	*80*	80	30	10	12	10	12	20
2300	5:00	6:00	*80*	80	30	12	12	15	12	20

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next higher band for possible openings.

bands will be shorter, occur closer to local noon, and provide paths mainly to the southern hemisphere with a possibility of transequatorial openings.

Twenty, thirty, and forty meters are both day and night bands. Twenty is the maximum usable band for DX in the northern directions these days during the daytime, then teams up with 30 meters to fill in through the night for the day-only bands. Forty meters becomes the main over-the-pole DX daytime band, with some hours covered by 30. This path may be affected by anomalous absorption during a few days of the month.

Eighty and one-sixty meters, the night-only DX bands, will exhibit short-skip propagation during daylight hours, then lengthen for DX at dusk. These bands follow the darkness path, opening to the east just before your sunset, swinging more to the north-south near midnight, and ending up in the Pacific areas during the hour or so before dawn. Eighty is the maximum usable band for some night hours now during sunspot minimum; consequently, signal strength and signal quality can be expected to improve. One-sixty may also be better. Remember the DX windows of 3790-3800, 1825-1830, and 1850-1855 kHz.

references

1. Garth Stonehocker, K0RYW, "DX Forecaster," *ham radio*, Decermber, 1984, page 63.
2. Garth Stonehocker, K0RYW, "DX Forecaster," *ham radio*, October, 1983, page 92.

ham radio

a scanner for
CB to 10-meter conversions

Diode matrix replaces 40-position selector switch

There have been many articles on CB-to-10 meter conversion projects over the last few years, one of which was an excellent article by MacFarquhar and Grant[1] in which they described the steps necessary to convert a Citizens Band AM radio to an FM transceiver operating on 10 meters. In this article, we show how to replace their 40-position channel selector switch with some digital logic that adds the ability to scan up to 20 channels. In addition, we described a way to obtain a 100 kHz offset for repeater work without using a second crystal. All the other features of MacFarquhar and Grant's original modification are preserved.

If you already have a converted CB, enough information is provided here to build the scanner; if you're starting the project from scratch, the MacFarquhar and Grant article will be necessary.*

circuit description

Since the CB was designed for mobile operation, the converted radio must use a mobile power source or operate from a 13.8 volt DC power supply. Power supplies are available commercially at surplus stores, and many excellent articles on building your own have been published.

Rather than build a separate +5 volt supply for the scanner, we borrowed a few mA from the +5 volt supply in the CB. Low-power components were selected to minimize the load on the existing CB radio power supply.

The scanner and control circuit requires only six integrated circuits and can be built on a 3 × 3 inch (7.6 × 7.6 cm) card. A diode encoder is used to program the operating frequencies. This gives the advantage of being able to select any of ten frequencies and more importantly, gives one a choice of crystals (frequency) to use.

A schematic diagram of the scanner section is shown in **fig. 1**. U1 is a decade counter with a built-in binary to decimal decoder. The outputs feed the diode encoder and indicator LEDs. The input to U1 comes from oscillator U2 (in the scan mode) or flip-flop U3A (in the single step mode). The other half of U3 is used to offset the frequency by 10 kHz. The frequency of the original CB is controlled by a Phase Locked Loop integrated circuit, or PLL IC. The diode encoder converts the ten decimal outputs from U1 to four control inputs to the PLL. The four control inputs cause the PLL to change its output frequency and the receive/transmit frequency of the radio. U4A is used to monitor the squelch of the radio and stop the scan when a signal breaks the squelch.

A type 555 oscillator (U2) is used to provide a scan step to the counter. The 7555 is a CMOS version of the 555 timer; its operation is identical except that the 7555 draws much less current. Radio Shack carries a TLC-555 which is also low current.

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*Reprints are available from *ham radio*, Greenville, New Hampshire 03048, for $3.00 each.

The oscillator is cut off by grounding pin 4 or 5. Pin 5 is connected to S3, the SCAN/LOCK switch. Pin 4 is controlled by U4A, a comparator. When the radio receives a signal that breaks the squelch, the voltage at Q120 collector goes up, forcing the output of U4A low, stopping the oscillator. The other half of U4 is used to drive a "busy" light for visual indication when a signal is present.

The diodes at the input to the counter form an OR circuit so that inputs to either diode will step the counter. U3A is a debouncer for the STEP switch, which should be a push button or spring loaded toggle. U3A is used on an R-S flip-flop — even though it is a D-type device — by wiring the switch contacts directly into the Reset and Set inputs.

If S2 is closed, the output of U3B will be held in RESET state, forcing a logic zero to PLL control pin P0. Frequency coverage will be ten "even" frequencies — 29.50, 29.52 through 29.68 MHz. If S2 is open, U3B will be toggled by the end carry from the counter each time the counter counts ten pulses or wraps around. When the output of U3B is high, frequency coverage is shifted up 10 kHz to cover the "odd" frequencies of 29.51, 29.53 through 29.69. The circuit will alternately scan the even then odd channels while S2 is open. When the frequency is shifted up, the high level at U3 pin 9 turns Q11 on, lighting the +10 kHz LED.

The ten outputs from the CD4017 counter drive 10 LEDS in addition to feeding the encoder. Q1 through Q10 (and Q11) can be almost any general purpose NPN small signal transistor such as a 2N2222 or 2N3904. The 4.7 kilohm base resistors were chosen to provide about 1 mA of base current, which allows practically any transistor to be used. Note that there is only one dropping resistor (1K) used for the ten LEDs since only one LED is on at a time. The current for the LEDs comes from the 13.8 volt source.

A diode encoder is used to convert from a decimal number to the digital data required by the CB PLL chip to select different frequencies. **Figure 2** shows the encoder used with our selected output frequencies and crystal frequency.

Figure 3 shows a transmit offset circuit that can be added to obtain the −100 kHz offset required for repeater access. Two boards were built using the crystal switching scheme described by MacFarquhar and Grant. They worked fine, but the price of the 10.795 MHz crystal exceeded the price of the CB board. A simple circuit was therefore developed to make use of the existing counter U1 when working through a repeater. The circuit (fig. 3) consists of two cascaded timer circuits, where U5 is a one-shot enable for oscillator U6. When the microphone PTT switch is depressed, the voltage on the CB point 14 goes positive and is inverted by Q13. The resulting negative shift at the collector of Q13 is coupled via the 470 pF capacitor to a diode gate and to the input of U5. The negative spike at pin 2 triggers U5, allowing oscillator U6 to send five pulses to the CD4017 clock input via the diode, which becomes the third leg of the existing OR circuit. When the PTT switch is released, the negative going level change at CB point 14 is coupled through the 0.01 microfarad capacitor to the input of U5 and another string of five pulses is sent to the 4017 counter. The circuit is disabled for simplex operation by grounding pin 4 of U5.

The pulse rate of U6 is about 240 microseconds. It is enabled for about 1 millisecond by U5. The time that U5 enables U6 must be set carefully. When the entire scanner is operating, the pulse width of U5 can be trimmed by selecting the proper value for R_X. Use a potentiometer or a decade box and lower the value of R_X until four pulses are output from U6. The pulses can be counted by using the PTT switch and observing the LEDs. Note the value at which the pulses change from five to four. Raise the value of R_X until six pulses are observed. Note this value and install a fixed resistor for R_X that is about halfway between these two values. If the 0.005 μF capacitor is accurate R_X should be around 150 kilohm. A mylar capacitor should be used in this circuit.

frequency and crystal selection

The following explanation is provided to explain frequency control and how to use an available crystal. The CB Phase Locked Loop (PLL) was designed to operate in the range of 2.24 to 2.68 MHz. Attempting to operate the PLL outside this design range may require modification of the PLL lowpass filter.[3,4]

The present ARRL 10-meter FM band plan covers the range of 29.50 to 29.70 MHz. The national calling frequency is 29.60 MHz (similar to 146.52 on 2 meters). Repeater outputs are on 29.62, 29.64, 29.66, and 29.68, with their inputs 100 kHz lower on 29.52 through 29.58. 29.50 is used as another simplex frequency. The CD4017 decade counter forms the heart of the scanner and gives us access to all ten frequencies. The ten additional frequencies are obtained by using a flip-flop to program the low order bit on the CB's PLL. The ten "odd" frequencies are presently all simplex frequencies.

To see how a crystal is chosen, we start at the required output frequency and work backward. If, for example, 29.6 MHz is the output frequency selected, the PLL VCO is mixed with a 10.695 MHz crystal oscillator to generate the transmit frequency. Our 29.6 MHz target requires that the VCO operates at 29.60 + 10.695 or 40.295 MHz. This in turn is mixed with the third harmonic of another crystal oscillator (Q105 on

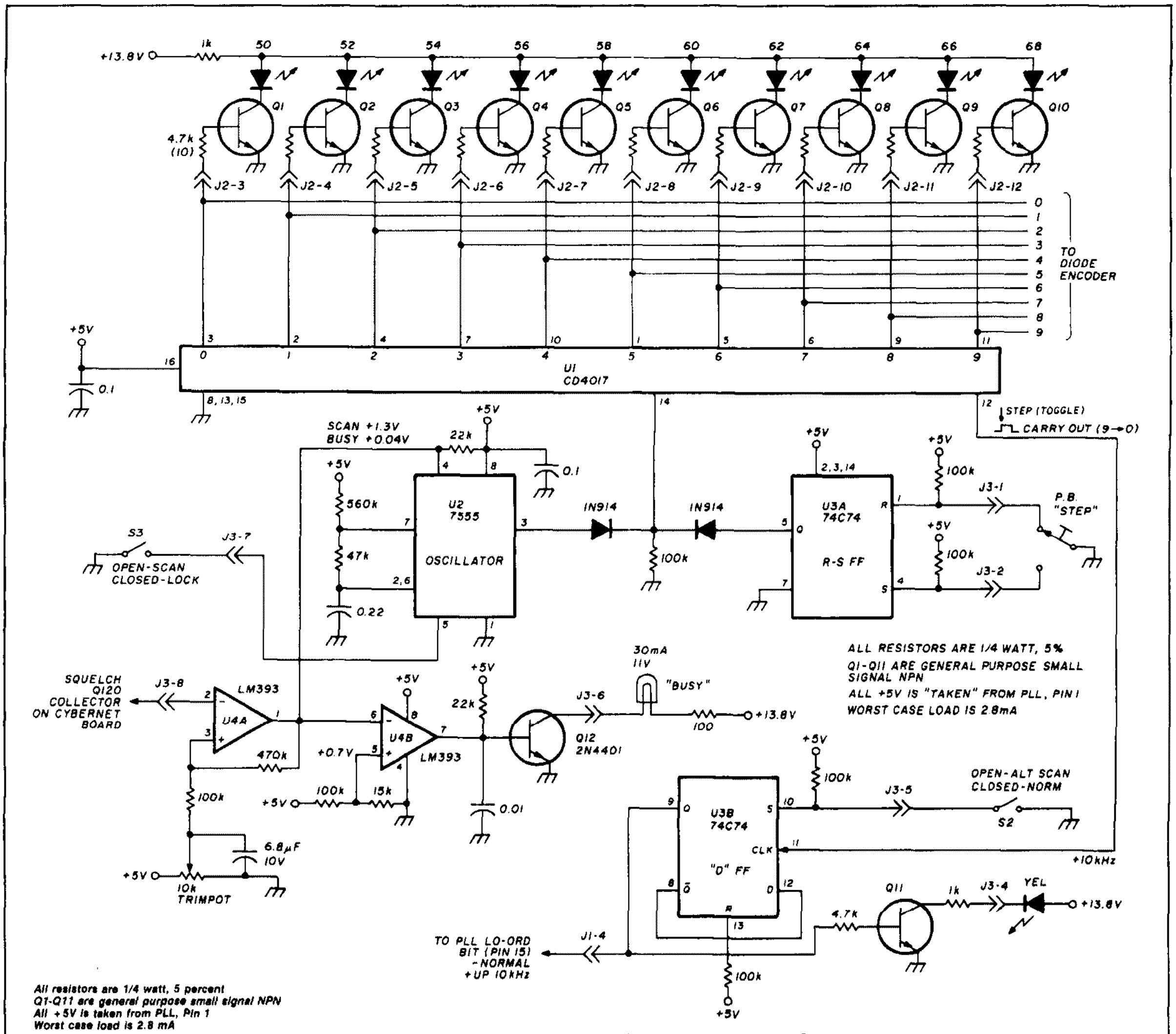


fig. 1. The scanner section only requires a few ICs.

the CB). The difference frequency is filtered and applied to the PLL. This frequency must fall in the range of 2.24 to 2.68 MHz.

When we started this project, we had about a dozen crystals left over from an earlier 10-meter club project. These crystals were marked 12.61333 MHz and had been ordered for a PLL using a reference frequency of 5 kHz rather than 10 kHz. The scheme we used required the crystal to operate up or down one third of a 5 kHz step or 1.67 kHz. We found the oscillator (Q105) more agreeable to being "rubbered" up than down, and had no trouble reaching 12.615 MHz. This frequency would have resulted in having to shift down to 10 kHz from 29.7 to reach 29.69 and not being able to use 29.50, which was clearly not desirable. By reducing C118 to 15 pF we were able to move the frequency up to 12.61833 MHz. Two units were built with crystals of 12.61833 MHz and they worked fine with the original value of C118.

The crystal's third harmonic of 37.855 MHz, when subtracted from the VCO frequency of 40.295 MHz, yields a difference of 2.44 MHz. To match at the 10 kHz PLL reference frequency, the PLL divider must divide by 244. The divide number is controlled by applying logic levels to the P0-P8 inputs of the PLL. Note that the PLL chip has internal pull-down resistors on these inputs. An open circuit is a logic zero, and a logic 1 is obtained by pulling an input up toward +5 volts.

We can develop a truth table based upon our choice of crystal. The P inputs of the PLL have binary "weights" with PØ having a value of 1, P1 having a

table 1. Truth table.

desired frequency (MHz)	VCO F (MHz)	mixer F* (MHz)	divider**	PLL program pin values P8	P7	P6	P5	P4	P3	P2	P1	P0
29.50	40.195	2.3400	234	0	1	1	1	0	1	0	1	0
29.51	40.205	2.3500	235	0	1	1	1	0	1	0	1	1
29.52	40.215	2.3600	236	0	1	1	1	0	1	1	0	0
29.53	40.225	2.3700	237	0	1	1	1	0	1	1	0	1
29.54	40.235	2.3800	238	0	1	1	1	0	1	1	1	0
29.55	40.245	2.3900	239	0	1	1	1	0	1	1	1	1
29.56	40.255	2.4000	240	0	1	1	1	1	0	0	0	0
29.57	40.265	2.4100	241	0	1	1	1	1	0	0	0	1
29.58	40.275	2.4200	242	0	1	1	1	1	0	0	1	0
29.59	40.285	2.4300	243	0	1	1	1	1	0	0	1	1
29.60	40.295	2.4400	244	0	1	1	1	1	0	1	0	0
29.61	40.305	2.4500	245	0	1	1	1	1	0	1	0	1
29.62	40.315	2.4600	246	0	1	1	1	1	0	1	1	0
29.63	40.325	2.4700	247	0	1	1	1	1	0	1	1	1
29.64	40.335	2.4800	248	0	1	1	1	1	1	0	0	0
29.65	40.345	2.4900	249	0	1	1	1	1	1	0	0	1
29.66	40.355	2.5000	250	0	1	1	1	1	1	0	1	0
29.67	40.365	2.5100	251	0	1	1	1	1	1	0	1	1
29.68	40.375	2.5200	252	0	1	1	1	1	1	1	0	0
29.69	40.385	2.5300	253	0	1	1	1	1	1	1	0	1

1 = logic one (+5)
0 = logic zero (ground)

Note 1: Q105 oscillator 12.61833 MHz new crystal.
Note 2: Oscillator × 3 = 37.85499 MHz.
*2.680 maximum, 2.240 minimum.

$$^{**}divider = \frac{(F_{OUT} + 10.695) - (3\ oscillator)\ MHz}{10\ MHz}$$

value of 2 and so on up to P8 with a value of 256. The frequency choice of 29.50 to 29.68 MHz works out so that P8 is always zero (floating), and the P5, P6, and P7 inputs are always logic 1, which we tied up to +5. To cover the desired frequencies, we need control only four inputs, P1 through P4. Since we have shown that the "divide by" number for 29.60 MHz is 244 and we are spacing our steps 20 kHz, it follows that 29.62 MHz would require two 10 kHz counts higher on 246. From the information that we now have, the truth table shown in **table 1** has been developed.

Examination of the table reveals that the P0 input is a logic zero for all of the "standard" 10-meter FM channels, which happen to be represented by even numbers. Switching the P0 input to 1 (+) shifts the PLL up 10 kHz to ten "in between" frequencies which are odd numbers — 29.51, 29.53, 29.55, etc.

Table 1 also shows the frequencies at the VCO output and the mixer output. Note that the mixer frequency must fall in the range of 2.24 to 2.68 MHz for the PLL to operate correctly. **Table 1** was derived from a spreadsheet program that does the mathematics involved with the crystal selection and frequency generation.† To check on the usability of a crystal, the user enters the crystal frequency and the highest output frequency desired; the program calculates the VCO output frequency, the mixer output frequency, the divider necessary to match the 10 kHz input to the PLL, and the PLL program pin values. The spread sheet is useful for demonstrating what is happening in the circuit as well as for crystal and frequency selection.

To build the encoder, replace each + in the truth table with a diode. No diodes are needed for the − points on the table because of the pull-down resistors in the PLL.

construction

The first step in building this radio should be to get the CB working as a CB. (The referenced articles cover this subject in detail.[2,3,4]) We took the additional step of inserting pins in the CB at the points where connections are to be made such as the speaker output, microphone input, and the volume and squelch potentiometers.

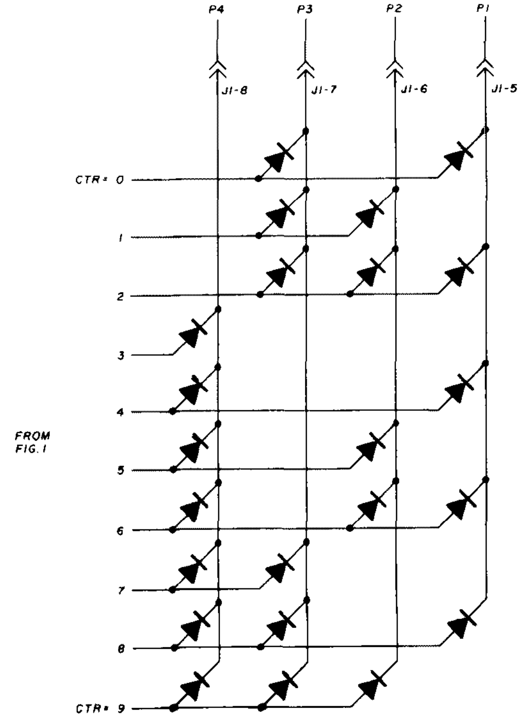
In addition to making the board easy to remove, we can interchange boards when problems arise. We found that the extra work was beneficial.

†A copy of the spread sheet, written in Lotus 123 and VisiCalc for an IBM PC and compatibles, or in VisiCalc for Apples, is available from the authors. Send an SASE for information.

The board used for the modification was made by Cybernet when the CB craze was in full swing. Cybernet manufactured these boards for several manufacturers such as Hy-Gain, Kraco, and others.

We found some CB boards with wrong value components. If the phase locked loop can not be made to lock by adjusting the VCO coil slug, check the value of R103 in the VCO circuit. The correct value is 1.5K, not 3.3K. Also, C101, which is buried in wax, should be 18 pF. (Some had a purple mark and measured 22 pF. If C101 is black, it's OK. This is apparently a production change that works on 27 MHz but doesn't work on the high end of 10 meters.)

We had a severe birdie problem on all 20 channels during tune-up of one of the first conversions that we built. The birdie was apparently caused by the 10.695 offset oscillator's being off in frequency. To prevent problems when the radio is being tuned up, all three oscillators must operate at the correct frequency. Radio Shack carries 3-10 pF trimmers (catalog No. 272-1338) that fit directly in the board; they are useful for adjusting the oscillators. Start with the 10.24 MHz reference oscillator. Loosely couple a frequency counter to the collector of transistor Q104. If the frequency is off, install a trimmer in location CT103 and try to move the frequency. C178 may have to be changed from 56 pF to a 39 or 43 pF capacitor. Set the frequency to exactly 10.24 MHz. To adjust the 10.695 MHz offset oscillator, connect the frequency counter to the emitter of Q109. Since the oscillator is not buffered, any appreciable capacitive loading will pull the oscillator frequency. We used a 10 to 1, low capacity oscilloscope probe to connect the counter. Install a trimmer in location CT102, and change C127 to a lower value. Adjust the CT102 for 10.695 MHz out with the transmitter keyed. Next connect the frequency counter to the dummy load at the output. Set up the P0-P8 PLL inputs to a known value such as 29.60 MHz and tune CT101 for the correct output, again with the transmitter keyed. If C118 must be changed, use the largest value that will permit CT101 to bring the oscillator to the desired frequency. This will make CT101 much less critical to adjust.



binary value	"P" No.	PLL pin No.	J1 pin
(16)	P4	11	8
(8)	P3	12	7
(4)	P2	13	6
(2)	P1	14	5
(1)	*P0	15	4

*Controlled by U3B on fig. 1.
Note: Inputs P5, P6, P7, and P8 are tied to Ground OR +5 for a combined weight of 224. See text for details.
All diodes (23) are 1N914 or equal.

fig. 2. The diode encoder converts a decimal number to the digital data required by the CB PLL chip.

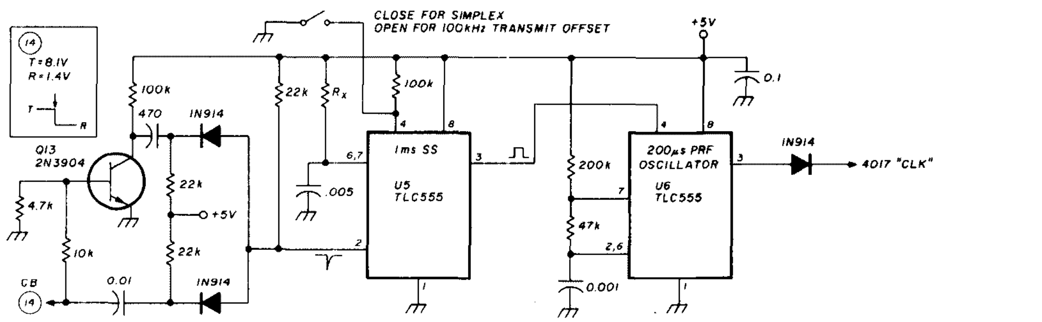


fig. 3. Digital 100 kHz transmit offset circuit.

conclusion

The scanner approach described above adds a new dimension to the CB-10 meter conversion by providing a simple yet versatile circuit. The digit offset control also has the capability of shifting up 100 kHz as well as down, in case you want to use the rig as a repeater — one local Amateur (Bob, W2XL) answered a call in the repeat mode and managed to work a station in Mexico.

In addition to the communication possibilities of a low power 10-meter FM rig, many local Amateurs use the radio to monitor 10 meters for isolated band openings. A dedicated rig eliminates the need to change frequency to check conditions; the radio is set up to monitor several repeaters. With the sunspot cycle on the decline, 10 meter openings are more sporadic, and when they occur, it's nice to know about them.

The circuits covered in this article can be used as shown or they can be used as a starting point for other innovative designs. There are other ways of obtaining the correct frequencies using a scanner approach such as using a decade counter and an adder to go directly into the PLL, which would make it easy to add the repeater offset. That may be a subject for a future article.

references

1. Ian MacFarquhar, VE3AQN, and Ken Grant, VE3FIT, "CB to 10 FM — One Group's Approach," *ham radio*, February, 1983, page 16.
2. Bob Heil, K9EID, "Experience 10-Meter FM Operation," *QST*, August, 1981.
3. J. Robert Witmer, W3RW, "Modifying a CB-Board Synthesizer for Amateur Use," *QST*, March, 1983.
4. Penn Clower, W1BG, "CB to CW? — Converting the HY-Gain Board," *73*, July, 1982.

ham radio

external product detector improves receiver performance

Add this circuit to your R390A or other older receivers

This unit is a combination of an IF amplifier, high stability local oscillator, wide dynamic range mixer and a separate audio amplifier circuit for SSB and CW reception. A stable high-current power supply is included so that the user needs only a speaker, a length of RG 58/U coax and a 120 VAC source of power. There is no need to re-enter the receiver past the IF output stage. The unit itself (fig. 1) is housed in an aluminum shielded cabinet.

Assembly and testing are straightforward. The only special equipment needed is a digital counter to set the oscilloscope and a 1 MHz heterodyne oscillator to the required frequency.

The construction and testing of each circuit should be done in the following order, completing each element before proceeding to the next:

- power supply
- audio amplifier
- VFO unit
- mixer-audio preamp (combined on board 3)

The power supply is a typical textbook assembly, with a three-terminal regulator. The bleeder R14 is used so that a constant load is placed on the power output at all times. To set the voltage, *before* AC is applied to the transformer, set R13 to the mid-point of the trimmer. Attach a voltmeter to the output point (at R14), apply 120 VAC to the primary of the transformer, observe the voltage at R14 and adjust trimmer potentiometer R13 to 12 volts. This circuit will need no further adjustments.

The assembly of the audio section has several important considerations. The TDA 2002 or TDA 2003 must have a 5-square-inch (12.7-cm^2) heat sink. A piece of aluminum 2 inches (5 cm) per side and 2-1/2 inches (6.35 cm) long was used to sink the device. The sink can be grounded to the B minus (and cabinet), providing a current return path for power and an adequate sink for device heat. Each module of the working model was assembled on a 3 × 4 inch (7.6 × 10.6 cm) vector board, with push-pin mounting for components and sockets for the transistors.

Keep all leads as short as possible. The 5 PC leads were soldered to push pins staggered in alternate rows. The assembled module was speaker tested with an audio oscillator and an oscilloscope with satisfactory results. The low-frequency cut-off was about 100 Hz and the upper range fell off after 6 kHz.

The local oscillator was assembled on a separate vector board using a standard Colpitts design. Note that the oscillator/mixer portions are FET devices, and consequently are high impedance circuits. To prevent loading down, the circuit testing was done with 10X scope (probe) leads.

frequency adjustments

To set the center frequency to 455 kHz, first set the C2 capacitor to its mid-point. Set C1, the compression trimmer, also to about its mid-point. Measure the oscillator frequency at the output end of C7 with the counter set to 1 MHz range. The coil, L1, should have a tuned resonance of 455 kHz ± 50 kHz. Adjust the slug in L1 to a point well before it extends beyond the wind-

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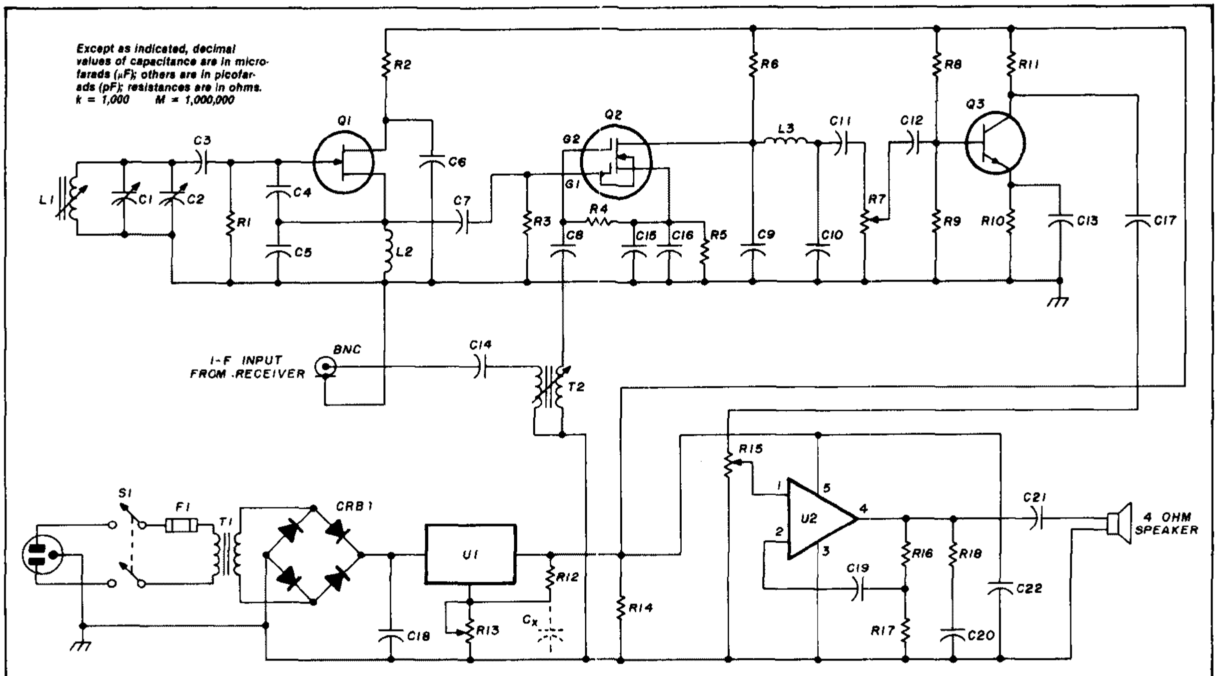

fig. 1. External product detector schematic.

Parts list.

item	description
BNC	BNC male, chassis mounting
C1	20-100 pF trimmer
C2	5-20 pF SFL ("butterfly") variable USB-LSB offset tuning
C3	390 pF, mica, 500 volt
C4	300 pF, mica, 500 volt
C5	500 pF, mica, 500 volt
C6	10 kilohm pF, ceramic, 50 volt
C7	100 pF, ceramic, 50 volt
C8	0.01 μF, ceramic, 50 volt
C9	1000 pF, ceramic, 50 volt
C10	5000 pF, ceramic, 50 volt
C11	1 μF, electrolytic, 50 volt
C12	1 μF, electrolytic, 50 volt
C13	50 μF, electrolytic, 10 volt
C14	0.01 μF, ceramic, 100 volt
C15	10 pF, 10 volt ceramic
C16	10 kilohm pF, ceramic, 50 volt
C17	1μF, electrolytic, 25 volt
C18	4700 μF, electrolytic, 50 volt
C19	470 μF, electrolytic, 100 volt
C20	0.1 μF, electrolytic, 50 volt
C21	1000 μF, electrolytic, 50 volt
C22	0.1 μF, electrolytic, 25 volt
CRB1	MDA970-1 Motorola
Cx	5 μF, 25 volt if the leads from regulator causes oscillation as seen on an oscilloscope
F1	buss fuse, 1 amp, 120 VAC
L1	J.W. Miller No. 23A474RPC
L2	1.2 mH, Miller No. 73F123AF
L3	1.2 mH, Miller No. 73F123AF
Q1	2N5486 Motorola
Q2	3N201 Motorola
Q3	2N2222 Motorola
R1	100 kilohm, 1/2 watt
R2	100 ohm, 1/2 watt
R3	100 kilohm, 1/2 watt
R4	10 kilohm, 1/2 watt
R5	100 ohm, 1/2 watt
R6	1.5 kilohm, 1/2 watt
R7	100 kilohm trimmer, 1/2 watt
R8	91 kilohm, 1/2 watt
R9	22 kilohm, 1/2 watt
R10	150 ohm, 1/2 watt
R11	3.9 kilohm, 1/2 watt
R12	270 ohm, 1/2 watt
R13	5 kilohm trimmer, 1/2 watt
R14	2 kilohm W.W., 5 watts
R15	100 kilohm potentiometer, 1/2 watt
R16	220 ohm, 1/2 watt
R17	2.2 ohm, 1/2 watt
R18	1 ohm, 1/2 watt
S1	DPST toggle switch
T1	115 volt primary, 17 volt, 2 ampere secondary
T2	455 kHz Miller No. 8812
U1	LM317K National

Note: All resistors 5 percent

ings, then fine tune by adjusting C1, while keeping C2 at its mid-range value. If leads have been kept short, the 455 kHz point should be found without difficulty. The open-air maximum drift as measured on a counter was less than 100 Hz at the end of 1 hour.

The oscillator output voltage through C7 was about 1.5 volts peak-to-peak, which is the required amplitude. If the voltage is greater than 1.5 volts, reduce the value of C7. The loading of this output by Q2 has negligible effect on the amplitude of the voltage.

A test for upper and lower sideband heterodyning is the most important aspect of a product detector circuitry.

With C2 set at its center point, the counter should still register 455 kHz. Adjust C2 slowly to a lower value of capacitance and frequency should increase to about 458 kHz. Move C2 to a higher capacitance (at the same excursion as above) and the frequency should drop to 452 kHz. Return the tuning to mid-point, 455 kHz. Turn off the power.

Install Q2 (3N201) in the socket. Connect a stable signal generator set to 455 kHz to the signal input BNC connector. The amplitude of the signal should not exceed 100 mV peak-to-peak sine wave, unmodulated. Connect the scope to the output end of C11, which is the top of trimpot R7. Power up the circuit and sync the VFO with the signal generator so as to display a null on the scope. Swing C2 through its upper and lower excursions, verifying that the waveform remains linear through sum and difference mixing of the signals. The amplitude should be about 200 mV peak-to-peak (3-dB mixer gain). Move the scope probe to the output side of C17. Adjust R7 trimpot for an output of less than 500 mV peak-to-peak. There should be no distortion of the output signal. If all components have been wired correctly, the Pi-section filter C9, C10, and L3 will have removed all mixer by-products with just the product of the two oscillators displayed on the scope. Increasing the signal generator output to 250 mV peak-to-peak should not cause distortion of the output signal. Be aware that the voice products of a product detector vary over the speech range and signal strengths. High signal amplitudes should not distort. A reduction of the R7 gain potentiometer will reduce large signal distortion.

The MOSFET, Q2, can tolerate high signal levels without distortion, however Q3 *can* be overdriven. A distorted signal is not pleasant to listen to.

After these tests have been completed, you're ready to assemble the power audio IC and hook up the R390A IF output connector.

The input and interconnect cabling was done with RG 217/U coax. The low frequencies of the system could have tolerated audio cords with equal success.

The builder can use edge card connect boards with matching sockets. This writer drilled holes in the vector board corners and fastened the boards with 4-40 screws and 1/2-inch (1.27 cm) aluminum spacers. It all works happily together and that's what it's all about.

modifying the R390A

Several changes can be made to improve sensitivity, and noise figure of the R390A receiver. The modifications are made to the IF module only.

Disconnect AC power to the R390A. The rear terminal board TB102 has a jumper between terminals 3 and 4. Disconnect this jumper and ground terminal 4 to the receiver frame. Connect a voltmeter between terminal 3 and ground. Carefully remove the aluminum shields covering T501, T502, T503, and Z503.

Drill 3/16 (0.476 cm) holes in the top center of these shield cans. Be sure to de-burr afterward since the aluminum is soft.

Inspect the IF transformer T501. Locate the resistors shunting the primary and secondary winding. *Caution:* The Litz wire connecting the inductors is very fragile and easily broken. The resistor values in my receiver were 47 kilohms but could be as low as 15 kilohms in other models. Clip out these resistors, and replace the shields and observe the AGC voltage. Switch the receiver on and set the function switch to AGC, note that the voltage measures approximately 28 volts in the 2 kHz selectivity position. When peaking T501, T502, T503, and Z503, the voltage should increase to 34 volts (minimum) if all tubes are working properly. Replace the jumper on the TB102 AGC line, connect the Product Detector Module and you'll discover that you have a hot SSB receiver that rivals anything built today.

using it in other receivers

The application of the Product Detector-audio assembly is not limited to the type of receivers discussed in the text. If the IF frequency is other than 455 kHz, (such as the R-388), then the local oscillator coil can be chosen to work with the required matching IF. Miller coils cover most of the wanted ranges; the builder can also wind a coil to the frequency of interest. The old Hallicrafters receivers, as well as the Hammarlund, National, and RCA of WW2 vintage can regain their vigor, providing the first local oscillator is stabilized. Don't scrap the old boat anchors just because they won't copy SSB or CW like some of the newer radios. Add this stand-alone module and find out how good it was all along!

The author has several R390A receivers in restorable condition, with the manuals. They are laboratory units, not military surplus. A QSL card to W6GB will get details. — Ed

ham radio

build a handy RF probe

Build it simple — or more complex, for increased sensitivity

An RF probe is a useful instrument to have around the shack — if you can find one that's sufficiently sensitive. This article will help you build your own, using solid-state Germanium or silicon diodes in simple circuits to provide the necessary sensitivity.

using diodes

Two principles must be understood in using diodes of all types: threshold of operation and detection sensitivity. Ideally, all diodes have a maximum sensitivity set by the diode equation:

$$I = I_S \{[exp. (qV/kT)] - 1\} \quad \textbf{(1)}$$

where I_S is the saturation current, which sets the threshold, q is the electron charge, k is Boltzman's constant, and T is the absolute temperature (room temperature is about 290 degrees absolute), and V is the voltage. The expression (q/kT) is extremely important in solid-state electronics and has a nominal value of 39. Its reciprocal is 0.026 volt.

The smaller the value of I_S, the larger the threshold voltage required with any of these devices.* With the germanium diode, the threshold will occur at an anode-to-cathode voltage in the range of 0.1 to 0.25 volt, and with the silicon diode, 0.5 to 0.75 volt.

In practice, an operating current level between 0.1 and 1 mA is selected to compare various devices. In normal operation, the anode-to-cathode voltage at which the chosen level of current flows controls the minimum sensitivity of a diode probe.

Measuring current as a function of voltage in solid-state diodes, one finds a two-to-one change in device current with approximately 0.018 volt change in applied voltage. This is the sensitivity needed (see appendix).

Using a diode having a low threshold and full incremental sensitivity, as it can be called can improve the minimum sensitivity of an RF probe to allow the measurement of voltages in the 100 to 500 millivolt range. (This is why hot-carrier diodes are important.) The simplest circuit for using a high-threshold-sensitivity diode is shown in **fig. 1**. The input capacitor is selected so that its reactance $(1/\omega C)$ is small compared to the diode's apparent resistance, and not more than the source impedance of the signal source. The resistance value in the output circuit should be large compared to the diode resistance, which is approximately $0.026/I$ where I is in amperes. Such an RF probe can be used in some receiver alignment applications as well as many routine tuning problems, and it can be used with a sensitive analog meter or a digital voltmeter.

building a better probe

Reducing dependence on the threshold limitation results in better RF probes because of high inherent sensitivity of solid-state devices. To take advantage of this sensitivity, a fixed level of operating current must be introduced to overcome the threshold and then apply a signal voltage from a low-impedance source to achieve the current sensitivity solid-state devices are theoretically capable of delivering.

Possibly the easiest way to avoid the threshold problem is to use a transistor so that we can sense the current changes more easily. A circuit using the LM 334 controlled-current source is shown in **fig. 2**. An isolation resistance that reduces the voltage drop across the LM 334 to perhaps a volt and a half will separate the DC and RF circuits.

*The value of I_S is a function of the numbers of the charge carriers in undoped or intrinsic material. In undoped germanium, there are 1131 times as many carriers than in silicon. This leads to at least 0.180 volt more bias for the silicon diode than for the germanium.

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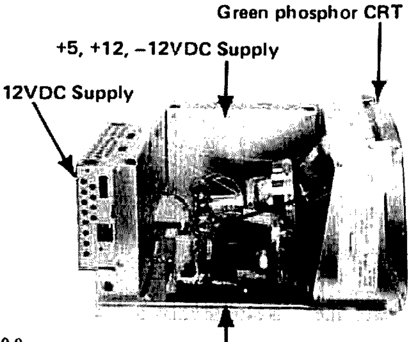

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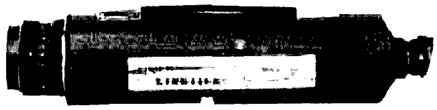

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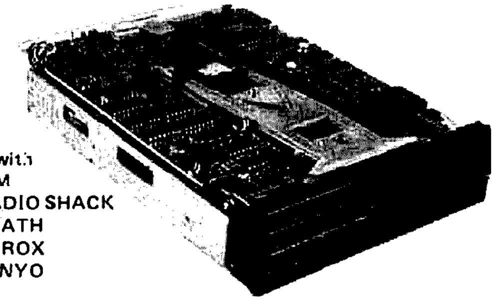

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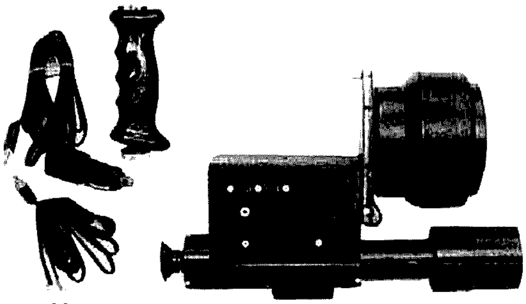

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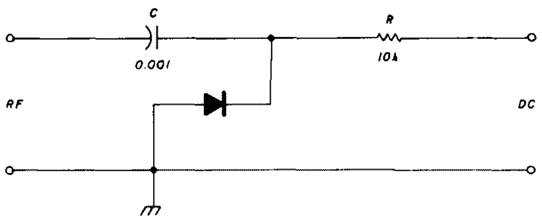

fig. 1. Simple probe. Note: resistor across RF terminals if needed.

A simple review of the principles underlying detection is included in the appendix.

The circuit shown in **fig. 2** works as predicted, but it still is not optimum because very small current changes must be observed if sensitivity is to be optimized. For this, a reference voltage follower can be used, (**fig. 3**), resulting in a voltage-sensitive bridge. Full-scale indication on the output meter with output changes of 100 millivolts (the full-scale reading of the meter) can be achieved easily. This may correspond to as little as 30 mV of RF signal. The scale will not be linear, but will approach a square-law function.

Using the meter directly between the collector circuit of the detecting transistor and the voltage reference will also reduce the sensitivity of the detector. Two circuits that can do this, both reducing the current load on the detector output, are shown in **fig. 4**.

The basic problems are now solved. To minimize the loading the detector places on the source, a situation that is particularly important if the RF source has high-impedance characteristics, VMOS transistors such as the VN10KM can be used as an amplifier. These transistors are better used as wideband amplifiers, with the output taken in the drain circuit. A small amount of degeneration through the use of a source resistance is acceptable, although it does reduce the amplification. In no case should the amplification exceed ten, and it may be as small as 1.0 to 1.5. A possible circuit is in **fig. 5**.

If greater sensitivity is required, a pair of high-frequency bipolar transistors can be used in a cascode arrangement after the VMOS transistor and before the input of the detection transistor to get an additional X10 voltage gain. A useful circuit for this is shown in **fig. 6**. This circuit should also be broadband. The output load resistance required will be approximately $260/i_C$, with collector current, i_C, measured in milliamperes. The resulting unit will have a voltage gain of roughly 10, and can be built quite easily. It will extend the minimum sensitivity down to about the millivolt level. These amplifiers limit with input signals over about 10 millivolts, and will also rectify, distorting the input signal.

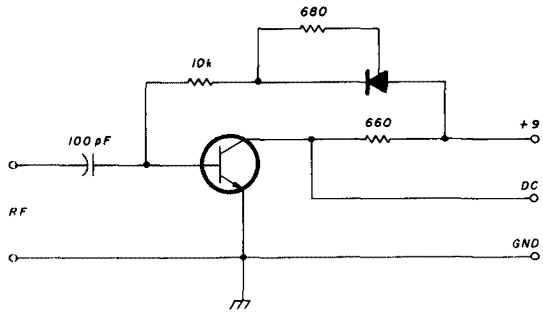

fig. 2. Sensitive probe using an LM334 controlled-current source.

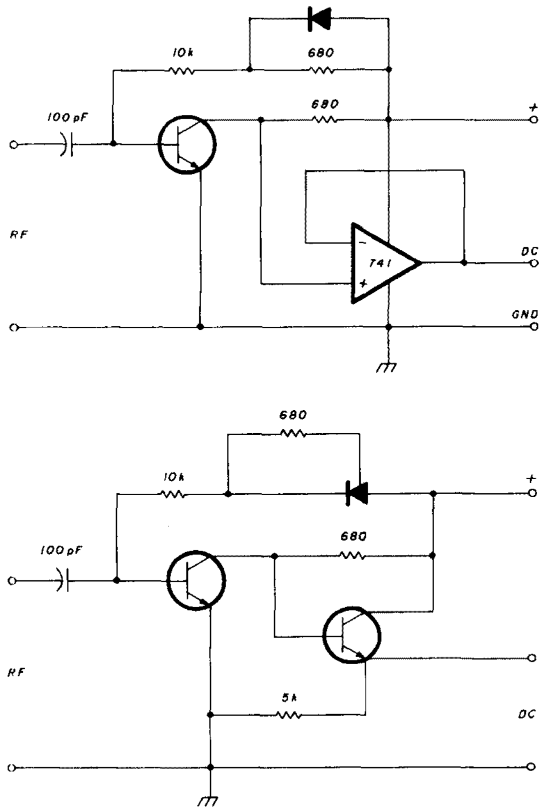

fig. 3. Two circuits that can be used to improve the sensitivity of the detector.

other applications

The buffered RF probe has a variety of applications of interest to Amateurs beyond its use in circuit testing. I have built one into a homebrewed Q meter, for example, and also into a low-frequency dip meter.

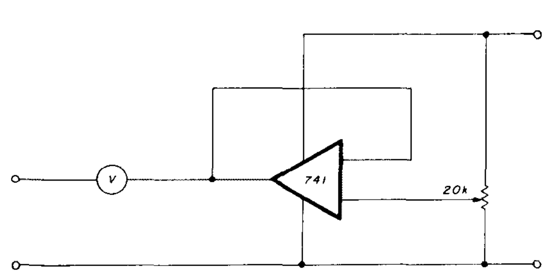

fig. 4. Circuit for a reference voltage follower to obtain a voltage-sensitive bridge.

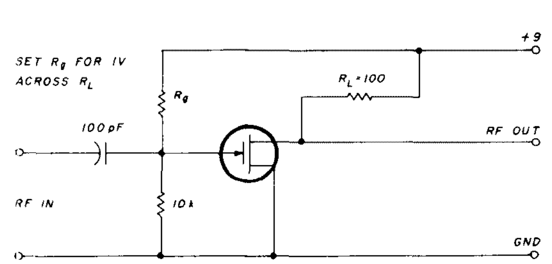

fig. 5. High-impedance preamplifier. Gain should not exceed 10 and may be as low as 1.0 to 1.5.

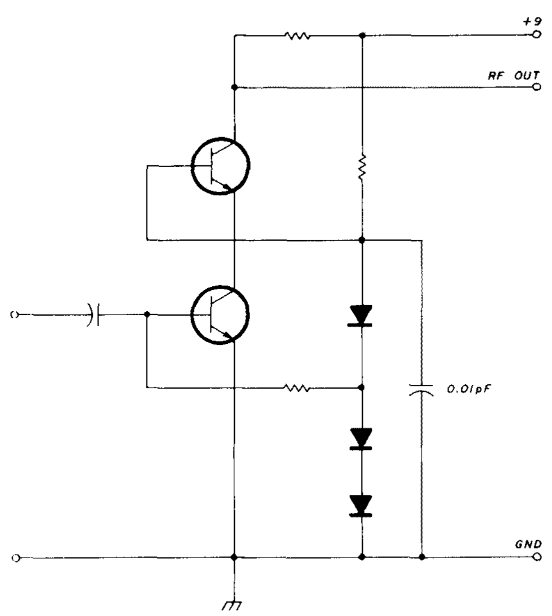

fig. 6. Cascode preamplifier with a voltage gain of about 10. Sensitivity is at the millivolt level.

The buffer FET can be the VN10KM mentioned above. (I have described an instrument for testing FETs in *Design of Transistor Circuits, with Experiments*.[1]

The field-intensity meter uses the cascode amplifier operating into the transistor metering circuit. I did not use the FET here because I assumed that the dipole antenna, consisting of two collapsible TV antennas, would have essentially a low-impedance output. (A suitable loop may be used instead, however.) The unit seems to work quite well.

The low-frequency dip meter must use variable-inductance tuning because a variable capacitor of appropriate size for this application would be much too large, and smaller ones would have too high reactance. The basic oscillator circuit I used is based on an emitter-coupled amplifier with a variable LM 334 current source for the base drive on both transistors, **fig. 7**. The oscillation level is set with a potentiometer across the LM 334 control points, and magnetic coupling is used to test a circuit. I have arranged the assembly so that I can plug in a variety of slug-tuned coils; the coils have slugs on threaded rods with a short piece of piano wire soldered in the screw slot. The drive consists of a piece of quarter-inch brass rod drilled and slotted to engage the wire. The frequency change is slow and smooth.

I can also tap off RF output from the FET buffer for frequency counting or use as a signal source. This unit easily fills the gap between audio and 2 MHz, the lower limit on most dip meters. The actual circuit is a combination of **figs. 3**, **4**, and **7**.

The Q meter used the same metering circuit as the dip meter for measuring RF. The RF source I use is a conventional inexpensive signal generator, except that I have tapped in on it prior to its normal output stage. These deliberately generate distorted waveforms to create harmonics in the output. You may wish to put an FET buffer between the oscillator and the special output to ensure better frequency stability. A suitable circuit for this is shown in **fig. 8**. About 10 millivolts of RF are needed by the Q meter input circuit.

The RF drive transistor that provides the RF signal to the measurement circuit should be a high-frequency bipolar transistor, and should provide a voltage gain of approximately 10 overall. (Only a hundredth of this is actually applied to the tuned circuit, 100 microvolts.) The full output is used to calibrate the circuit, and the part to control the tuned circuit. One is then matched against the other, based on oscilloscope calibration. A possible circuit is shown in **fig. 9**. Minimum Q sensitivity will be about 20. About 25 mA of collector current will be needed by the 2N2222 or 2N2369 to provide the required signal voltage. A Q test circuit is shown in **fig. 10**.

calibration

This calibration technique, done with an oscilloscope, can be used wherever minimum or maximum

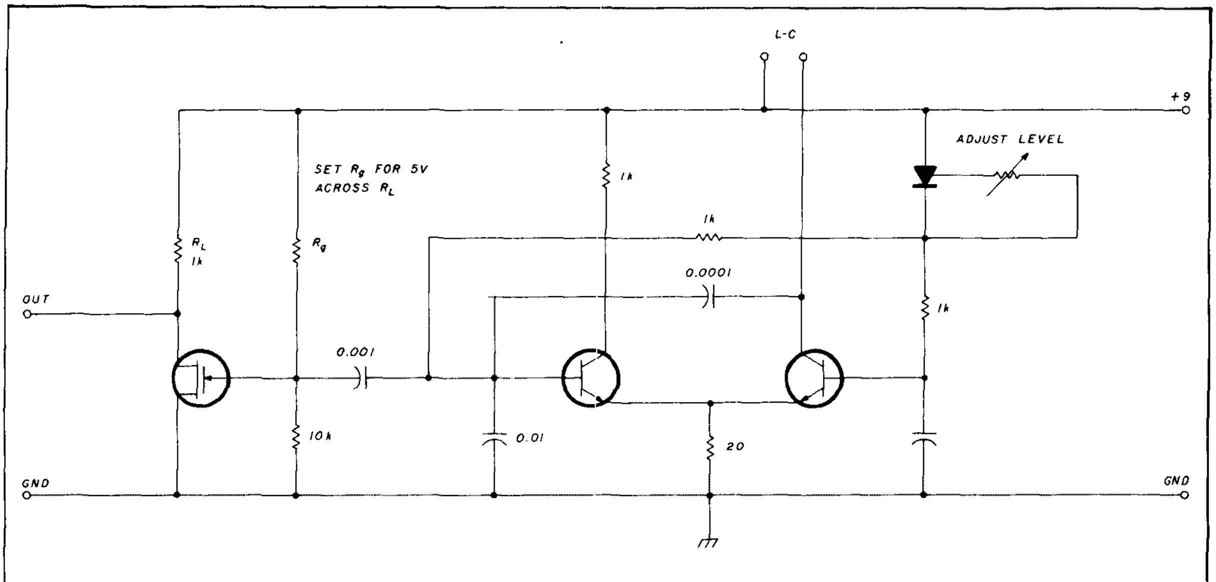

fig. 7. Oscillator for low-frequency dip meter.

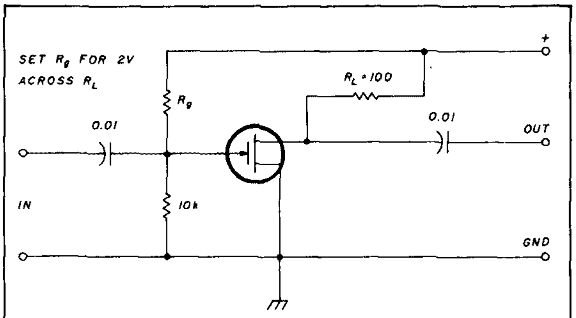

fig. 8. Buffer stage to ensure better frequency stability for Q meter use.

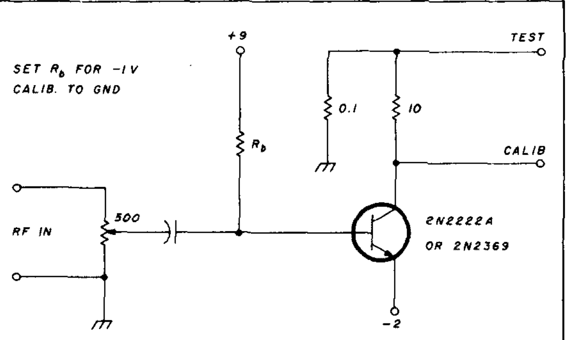

fig. 9. Q driver has a minimum sensitivity of about 20.

deflection is insufficient to assure proper circuit operation, but an actual voltage value is required. The oscilloscope can be calibrated with a reference voltmeter and an AC signal. RMS calibration should be used. The low-frequency calibration is used, with the help of the oscilloscope, to calibrate sinusoidal signal well within the upper limit of the oscilloscope (1/2 to 1 MHz for a 5 MHz unit). This calibration then is transferred to the Q meter measuring circuit. The source calibration voltage for setting the Q meter reference input may be adjusted for 20, 50, and 100 millivolts for the reference excitation. This will make possible measurements of Q to 100, 200, and 500. In addition, the meter is calibrated for a series of input signals ranging from about 10 mV to 100 mV, at least every 10 mV, to provide the actual Q measurement. A special scale may be made for the meter if desired. Since the Q is proportional to the RF voltage generated, calibration is important.

concluding remarks

These simple probe circuits can easily be built into transmitters, and receivers in which critical tuning is required as well as into other useful test instruments. In all cases, it is important to minimize probe loading on the circuit being tested without simultaneously degrading sensitivity. The VMOS circuits are ideal for this purpose, and they do not limit as readily as bipolar transistors. Where either germanium or silicon diodes can provide adequate sensitivity, they, of course, are the component of choice, but they do have the excessive threshold voltage requirement that the current-stabilized transistor circuits can help overcome.

RF measurements have always been a stumbling block in the typical Amateur station, including mine. I can remember building a nice wavemeter that I was

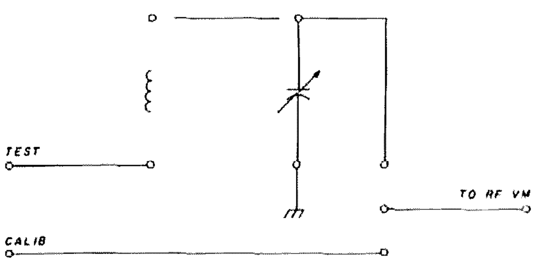

fig. 10. Q test circuit using an RF voltmeter. High-impedance input is used.

unable to calibrate. The described circuit techniques are a result of years of trying to solve similar problems inexpensively. Perhaps these ideas can help rekindle the experimental spark in other Amateurs, and help reverse the trend away from homebrewed equipment. *You don't have to be a mathematician or physicist to contribute to progress!*

reference

1. Keats A. Pullen, Jr., *Design of Transistor Circuits, with Experiments*, Howard W. Sams, Inc., Indianapolis, Indiana.

appendix

Rectification is a change in the effective resistance (of conductance) of a diode as a function of its applied voltage. This difference causes distortion, and it is the distortion of the current-voltage relationship that converts an AC signal into some level of change of average current and average voltage.

Strictly speaking, the relationship between the current and the voltage in a diode is nonlinear, or exponential. But for small voltages, it "looks" like a square-law device. The basic relationship can be stated in terms of the equation:

$$i = I_o \exp(qV/kT) = I_o[1 + qV/kT + (qV/kT)^2/2 + \ldots] \qquad \text{(A1)}$$

When V is set equal to $V_o \sin(2\pi ft)$, the relationship becomes, if $a = (q/kT)$:

$$i = I_o\{1 + aV_o \sin 2\pi ft + [a^2V_o^2 \sin^2(2\pi ft)]/2 + \ldots\} \qquad \text{(A2)}$$

The third term in the brackets on the right converts to the form:

$$(a^2V_o^2/4)(1 - \cos 4\pi ft)$$

and the DC term now is $(1 + a^2V_o^2/4)$. The term $a^2V_o^2/4$ represents the DC shift due to detection, and the current peaks in the two directions are:

$$I_o(1 \pm aV_o + a^2V_o^2/2) \qquad \text{(A3)}$$

The capacitor-input detector has a peak shift from $(1 + aV_o + a^2V_o^2/2 \times I_o$ to $(1 - aV_o + a^2V_o^2/2) \times I_o$. We filter out the $V_o \sin 2\pi ft$ and the $\cos 4\pi ft$ terms with the RC circuit. With the peak detector, the current change is $I_o(a^2V_o^2/2)$; this is what we wish to read.

From the above relationships, it is possible to show that with the *normal* diode detector, a current change of 2:1 can be obtained with 18 millivolts peak-to-peak signal on a diode, and a DC shift between 8 and 16 percent may be generated, depending on whether square-law or peak detection is selected. The signal distortion is roughly 8 percent.

ham radio

product REVIEWS

Butternut HF-2V vertical antenna

About seven years ago, Butternut burst on the scene with one of the first "new" antenna ideas for the Amateur market. Their HF-5V was a five-band, no-trap antenna that covered 80-10 meters. I was one of the first to get one for review and was very much impressed with its performance. In head-to-head competition with comparable antennas, the Butternut always was at least one S-unit stronger during on-the-air tests. Owners of Butternut's current model, the HF-6V, have told me that it does equally well.

Recognizing that the six-band HF-6V was a little short for 80 and 40 meters and the fact that the high bands are going to be at best marginal for the next few years, Butternut modified their basic antenna design to optimize performance on 80 and 40 meters. The net result is the HF-2V antenna. The HF-2V incorporates all of the design features of the HF-6V — stainless steel hardware, T-6061 aluminium tubing, for strength and a double-walled base section for even more structural integrity and no traps.

the antenna

As with any antenna project, the first task was to read the instruction manual from cover to cover and take a full inventory of all the parts and hardware. All you need to assemble the antenna is a blade screwdriver, pliers, and a knife. A set of nut drivers would also be handy.

Assembly is straightforward and should only take an hour or so to accomplish. Identification of parts and hardware is easy if you use the schematic diagram that comes with the unit. The antenna is constructed of telescoping aluminum with sections held together with stainless nuts and bolts.

The antenna is loaded four feet from the base with an L/C combination made of very heavy duty aluminum wire and high voltage transmitting mica caps. Since the antenna is almost a full quarter wave on 40, very little of the 40-meter coil is used to tune the antenna. Should there be too much inductance for 40, a shorting strap is provided. The antenna is impedance matched at the base with an adjustable coil.

about radials

A lot of fuss is made about radials for vertical antennas and rightfully so. Many of those who install verticals constantly complain that the performance is less than was expected. In a majority of the cases, the prime reason is that there were either no radials or an insignificant number of radials under the antenna. In Butternut's opinion, it's best to assume that the earth you're working over is poor and that you'll need to engineer an artificial ground of your own. Radials can be either buried or laid atop the ground, whichever is easier. In general, it's better to install more short radials above ground than to lay just a few long ones. 16 50-foot radials will work rather well in most installations. The *ideal* radial system would include 120 resonant radials, 1/4 wave on each band. Needless to say, that is beyond the capabilities of most of us.

checkout and adjustment

Tune-up is simply a matter of adjusting the two loading coils for minimum SWR and adjusting the base impedance matching coil to correctly match the coax. As mentioned before, because the antenna is almost a full 1/4 wave on 40, very little inductance is needed to tune the antenna. To set up and tune it, you need to first short out a full seven turns of the 40 meter coil.

Now it's simply a matter of tuning the antenna for the lowest SWR on first 80 and then 40 meters. The next procedure is to attach the base step-up transformer and match the feedline to the antenna. This is a little trickier than it sounds, and took me more than a few minutes to accomplish. In order to get a proper match, you either expand or compress the coil to increase or decrease the inductance needed to get a proper match. I found that no matter how much I changed the coil, I couldn't get a proper match. Sometimes drastic measures are necessary; I used one here, reasoning that no matter how much I expanded the coil, I still had too much inductance to match the antenna. So I cut the coil in half, reattached it, and again applied power. The result was that the antenna was nearly matched and required only one minor change to complete tune up.

optional accessories and top loading

Butternut offers a number of accessories you can use to tailor the HF-2V to your own personal needs. As most low band operators will know, top loading antennas is a very practical way to make short antennas perform better. The top loading kit gives you four 25-foot stranded "umbrella" wires and insulators. The "umbrella" is designed to improve the overall antenna performance on 80 meters (and 160 with optional TBR-160 160-meter base loading kit). It will also improve the bandwidth of the antenna to over 100 kHz on 80 and approximately 35 kHz on 160 meters. Butternut also has a 30 and 20-meter resonator kit that can be added to the antenna. Unfortunately, you can only add one or the other, but not both.

operation

I've had this antenna up now for several months and have used it on both 80 and 75 as

well as on 40 meters and have been quite impressed with its performance. On 80 the antenna has performed as well as my 1/4-wave sloper, and actually outperformed it in certain directions and conditions. When Ron Wright was on Tonga, A35EA, I was able to work him through a fair-sized pileup without too much difficulty. Not being a fan of 40 meters, my operating experience there is admittedly spotty. But this antenna has performed very well and I've been able to work just about everyone I called.

The Butternut HF-2V is the perfect antenna for those who want good low band performance without investing a lot of time and money installing a larger antenna. It's performance is comparable with a number of other antennas, and HF-2V owners will get years of excellent results with this antenna.

It's interesting to note that the recent DX'pedition by the Texas DX Society took a Butternut HF-2V to the island of Desescheo. Used with the optional TBR-160, it accounted for thousands of contacts on 80 and 160. The expedition members have reported that the antenna was a breeze to set up and performed flawlessly. I remember that during that operation, night after night, they came pounding through with perfectly Q-5 signals on 160.

specifications

weight	131 pounds/5.9kg
height	32 feet/9.75 meters
feedpoint impedance	50 ohms
VSWR at resonance	1.5/1 over a suitable ground
VSWR < 2/1	65 kHz on 80, full band on 40
power rating	2 kW PEP/1 kW CW (less with TBR-160)
wind no ice	80 MPH/125 kph when properly guyed

—de N1ACH

MFJ-204 antenna bridge

The MFJ-204 Antenna Bridge is designed to do three basic tasks: measure antenna impedance, measure resonant frequency, and to tune antenna tuners "off the air." It's a handy piece of test equipment that every ham should have. There are several reasons that an antenna bridge is a much more versatile tool than its cousin the noise bridge. The noise bridge is basically a wideband oscillator that can be used to roughly determine resonant frequency or impedance of an antenna. I say "roughly" because the wideband oscillator is difficult to use, is imprecise and can sometimes give misleading results. The antenna bridge, on the other hand, has a tunable oscillator that can be set to the precise frequency you want to work on. The MFJ 204 covers 1.6-2.5, 3-4, 7-11, and 13-30 MHz so you get full coverage of all HF Amateur Radio bands.

To measure antenna impedance, the first thing you do is set the oscillator to the frequency you want to work on. This is done by loosely coupling the antenna bridge to your receiver and listening for the beat note. You then connect the antenna bridge to the antenna to be measured and adjust the resistance control for a dip on the antenna bridge's meter. Once you get this dip, you flip the bridge over and read the resistance from the factory-calibrated chart on the back of the unit. On my 160 vertical antenna, I measured an impedance of 35 ohms, just about where it should be.

Using the antenna bridge to both measure resonant frequency and tune an antenna tuner is pretty much the same process. When used with an antenna tuner, the antenna bridge eliminates one of the most annoying aspects of the hobby – the ubiquitous and infamous "tuner-upper."

The MFJ 204 measures 2-1/2 × 2 × 7 inches and weighs just about a pound. It will run off a 9-volt battery or an optional 110-volt AC power supply. It's covered by the standard MFJ 12-month warranty and should give many years of good, reliable service.

I've since used the antenna bridge on several antenna projects and have enjoyed its utility. As long as I can keep the review unit, I think I'll let the noise bridge gather dust.

N1ACH

NEW products

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The Computer Patch Model CP-100 Interface is a complete terminal unit for Morse, Baudot, ASCII, and AMTOR. It will interface a computer running communications software via TTL levels (RS-232 optional) to your radio. With the optional current loop provisions, the CP-100 can be used with a mechanical teleprinter also. The tuning indicator is a ten-segment bargraph featuring discriminator-type operation which graphically shows selective-fading and is ideal for AMTOR use (tuning 'scope outputs are also

available). The CP-100 also features a front panel squelch control which inhibits data to the computer when no signal is present (thus preventing print when receiving noise). Front panel selec-

tion of 110 and below or 300 baud operation is provided along with an internal monitor speaker. The CP-100 is housed in an attractive metal case, minimizing any potential RF interference efforts.

For further information, contact Advanced Electronic Applications Inc., P. O. Box 2160, Lynnwood, Washington 98036.

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smart patch plus™

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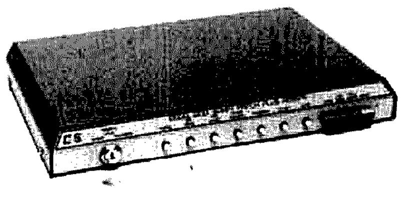

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AMATEUR RADIO CLASSES: Chelsea Civil Defense will sponsor evening classes at Chelsea High School starting October 16, 1985, for Novice (basic level ham license), and Tech/General licenses. Minimal cost for materials only. For information: Frank Masucci, 136 Grove Street, Chelsea, MA 02150.

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UHF PARTS. We stock GaAs Fets, trimmers, and many of those impossible to find parts for the UHF builder. For example: MGF 1202 at $10.00, Finger Stock at $7.50 per 16" strip. SASE brings list. Microwave Components, 11216 Cape Cod, Taylor, Mich. 48180

SELL: TS147A Test Set. 8.5-9.6 GHz CW-FM generator, frequency meter, power, manual. $50.00 plus UPS. Huntley, W6RNC, PO Box 478, Nevada City, Calif. 95959.

COMPUTERIZE YOUR IC-720 or R-70 Keyboard frequency entry, 64 memories, scanning (frequency, memory, and mode). Many more features. Requires no interface, just cable — directions included. For Commodore 64. Cassette $14.95, disk $16.75, ppd. Cables also available. David Oliver, W9ODK, Rt. 2, Box 246, Shevlin, MN 56676.

WANTED: Operations manual schematic for Sierra 305 A-L frequency selective level meter. Weber, 2605 West 82nd Place, Chicago, IL 60652.

WANTED: Old tube HiFi equipment. Working or not, complete or parts. All makes and models. Jack Smith, 59 Millpond, N. Andover, MA 01845.

RECEIVERS: Good operating and appearance with manual. HA-350, $50.00; NC-173; NC-300 $100.00 each. All plus shipping. Lisains, 116 Orton, W. Caldwell, NJ 07006 (201) 226-7943.

FREE 3 month subscription. Send callsign to Westlink Report-H11, 28221 Stanley Court, Canyon Country, CA 91351. New only!

KT5B ANTENNA 160m-10m no traps. $59.95. Weather Boot Kit $8.95. Open-wire, roller-inductor, antenna accessories and much more. Kilo-Tec, PO Box 1001, Oak View CA, 93022. Tel: 805-646-9645.

WANTED: R-1490/GRR-17 receiver, parts and manual. Gene Mottern, 313 Royal Drive, Kingsport, Tenn 37663.

2 METER AMP KITS: 8877 legal limit kit $395. 3CX800A7 900W kit $325. Also HV power supplies, CX600N relays, parts and EME newsletter. SASE for catalog. 2 Meter EME Bulletin, 417 Staudaher St., Bozeman, MT 59715

COMING EVENTS

Activities — "Places to go . . ."

GEORGIA: Southnet II Packet Radio Conference hosted by Georgia Tech Amateur Radio Club and sponsored by the Georgia Radio Amateur Packet Enthusiast Society. November 23 and 24, Georgia Tech campus downtown Atlanta. Tech sessions 8:30 to 5:00 Saturday and 8:30 to 12 Noon Sunday. For further information: Bill Crews, WB2CPV, 1421 Hampton Ridge Road, Norcross, GA 30093. (404) 923-1978.

MASSACHUSETTS: The Honeywell 1200 Radio Club and the Waltham Amateur Radio Association will hold their annual Amateur Radio and electronics auction, Saturday, November 23 at the Honeywell Plant, 300 Concord Road, Billerica. Doors open 10 AM. Free admission and parking. For more information: Doug Purdy, N1BUB, 3 Visco Road, Burlington, MA 01803.

ILLINOIS: Waukegan Squadron Civil Air Patrol will hold its annual Fall Hamfest, Sunday, November 3, Lake County Fairgrounds, Rt. 120 and 45, Grayslake. Doors open 7 AM. Admission $3.00. Tables $5.00. Large indoor area, cafeteria, free parking. For further information and reservations SASE to CAP, 637 Emerald St., Mundelein, IL 60060.

COLORADO: AMSAT will hold its third annual Space Symposium in Vail, November 9. AMSAT has issued a call for papers to be presented at the symposium. Topics may include those of interest to Amateur Radio Satellite enthusiasts. Abstracts may be sent to AMSAT, PO Box 8005, #281, Boulder, CO 80306. For further information: Molly Hardman, N3CHZ, Chairperson, (303) 939-9334 or AMSAT Headquarters (301) 589-6062.

MINNESOTA: The annual Handi-Ham Winter Hamfest, Saturday, December 7, Eagles Club, Faribault. Registrations 9 AM. Handi-ham auction of equipment, dinner at noon followed by a program. Talk in on 19/79. For information: Don Franz, WØFIT, 1114 Frank Avenue, Albert Lea, MN 56007.

NEW JERSEY: The 4th annual Jersey Shore Ham/Computer/Electronics Flea Market, November 3, 9 AM to 3 PM, Neptune City Recreational Hall, Neptune City Shopping Center, Neptune. Admission $3.00. Children under 12 and XYL's free. Refreshments available. Table $6.00. Tailgating $3.00. Send payment and SASE to Jersey Shore Hamfest, PO Box 192, West Long Branch, NJ 07764. Talk in on 147.045 and 146.52 simplex. (201) 222-3009.

LOUISIANA: The Twin City Hams are sponsoring a Hamfest, Saturday, November 9, West Monroe Convention Center, North 7th Street, West Monroe. 9 AM to 4 PM. Free swap tables. Amateur Radio exams will be given. Talk in on 146.52 simplex and 146.25/146.85. For more information: Benson Scott, AE5V, 107 Contempo Street, West Monroe, LA 71291.

MASSACHUSETTS: The MIT UHF Repeatur Association and the MIT Radio Society offer monthly Ham Exams, Novice through Extra, Wednesday, November 20, 7 PM, MIT room 1-134, 77 Mass Ave., Cambridge, MA. Reservations requested 2 days in advance. Contact Ron Hoffmann at (617) 253-5820/646-1641. Exam fee $4.00. Bring copy of current license, two forms of picture ID and completed form 610 available from FCC in Boston (223-6609).

WISCONSIN: The Milwaukee Repeater Club is proud to sponsor the "6.91 Friendly Fest" on Sunday, November 24, Serb Hall, 51st and Oklahoma. Bring your swapfest bargains and sample good food and beverages. Also bowling. Tickets $3.00. 4' tables $4.00. Save $1.00 — SASE with payment to Milwaukee Repeater Club, PO Box 2123, Milwaukee, WI 53201 before November 11. Talk in on 146.31/91 and 146.52.

ILLINOIS: The Chicago Amateur Club will hold its annual Ham Auction, November 6, 7 PM to 10 PM, Edgebrook Golf Course Field House, 6100 N. Central Ave., Chicago. For information: (312) 545-3622.

INDIANA: The Allen County Amateur Radio Technical Society's 13th annual Fort Wayne Hamfest, Sunday, November 10, 8 AM to 4 PM, Allen County Memorial Coliseum, Coliseum Blvd., US 30. Indoor tables available $8.00. AC power extra. Premium tables with AC $20.00 each. Admission $3.50 advance. $4.00/door. Children under 11 free. Ladies' activities, forums, banquet Saturday night. Nearby motels and restaurants. VE exams Saturday, November 9 advance registration only. Talk in on 146.28/88. For information or reservations: ACJ-ARTS Hamfest, PO Box 10342, Fort Wayne, IN 46851.

MICHIGAN: The Oak Park High School Electronics Club presents the 16th annual Swap N Shop, December 1, Oak Park High School, Oak Park, 8 AM to 4 PM. Admission $2.50. After 12 Noon $1.50. All 8' tables $8.00. SASE to Herman Gardner, Oak Park High School, 13701 Oak Park Blvd., Oak Park, MI 48237 or call (313) 968-2675.

OHIO: The Massillon ARC will sponsor "Auctionfest 85", November 24, Missillon K of C Hall, off Rt. 21, 8 AM to 5 PM. Sellers setup 7 AM. Admission $2.50 advance and $3.50 at the door. Tables available at $7.00 per 8' space. Refreshments and sit down dinner. Free parking. Auction 11 AM. Talk in on W8NP, 147.78/.18. For advance registration and information: MARC, PO Box 73, Massillon, OH 44646. SASE please.

NEW YORK: 7th annual "Ham-Central", Sunday, December 1, Main Hall, Lutheran School, Moriches Road, St. James, Long Island, close to NYC. Doors open 7:30 AM for sellers/dealers. 9 AM general public. Table space $10.00/8' includes two free admission tickets. General admission $3.00. XYL and kids under 12 free. Seminars plus ARRL and AMSAT. Food and refreshements available. Nearby shopping malls and historic sites. Talk in on 144.550/145.150 and 146.52 MHz. For reservations and information call after 6 PM Bob Yarmus, K2RGZ (515) 981-2709 or Andy Feldman, WB2FXN (516) 928-3868.

FIRST US YL WINS SWEDISH AWARD. Emily Maylan, AC2V, of Yonkers, NY, has become the first American YL to win the "100-SM" award for confirmed contacts with 100 Swedish hams. Only 659 other hams worldwide have previously won this award.

OPERATING EVENTS — "Things to do . . ."

The Armored Forces Amateur Radio Net (A.F.A.R.) will operate from 0000Z November 10 to 2400Z November 11 to commemorate Veterans' Day 1985. Phone 80, 40, 20, 15 meters. CW on 40 meters. Certificate available for contact with member station. Send #10 SASE to WB1DWR #90, 16 Berkeley Circle, Newington, CT 06111.

The Royal Jordanian Radio Amateur Society has announced plans for JY50, a two week celebration marking the 50th birthday of His Majesty, King Hussein, JY-1 starting November 7 through November 21. For the special "Worked JY50 certificate" send copies of log and either 10 IRC's or $5 (US) to the Royal Jordanian Radio Amateur Society, JY50 Celebration, PO Box 2353, Ammann, Jordan.

THE GUERRI REPORT

Ernie Guerri W6MGI

new components and techniques for RF designers

When manufacturers offer new equipment, the resulting design is a complex function of expected market, engineering hours, and the availability of necessary technology and components. Fortunately for designers of RF equipment, new components are rapidly being developed to keep pace with the boom in telecommunications.

Among the more exciting developments in the semiconductor industry is the emergence of commercial Gallium Arsenide (GaAs) ICs. Both analog and digital functions are being implemented. The digital functions — prescalers, counters, memories, and shift registers — are capable of operation at 2 to 3 GHz, allowing the design of phase locked loops and frequency synthesizers at low microwave frequencies using only a few chips. The required interstage amplifiers are now also available as off-the-shelf GaAs IC amplifier blocks (Microwave Monolithic Integrated Circuits, or MMICs) covering octave (2:1) bandwidths with 10 to 30 dB of gain. These gain blocks are available from 50 MHz to over 18 GHz. Prices are high right now but we can expect dramatic reductions as sales volume increases.

In the realm of discrete devices, the High Electron Mobility Transistor (HEMT) is just now ready to emerge commercially. This improved class of GaAs FET can provide noise figures of about 1 dB at almost 10 GHz, with 8 to 10 dB of gain! Several companies have developed low-cost ICs based on HEMT concepts which would integrate large portions of complex microwave receivers into a single chip.

One of the most fundamental concepts in modern telecommunications is frequency translation, or mixing. Most receivers and transceivers have several conversions from one frequency or band to another. An important characteristic of the translation device is its ability to minimize the generation of unwanted products. Mixer development has now advanced to the point where we have image rejection mixers (the image is automatically suppressed by about 20 dB) well into the microwave region. There is a similar class called "termination insensitive mixers" that perform well even with highly *reactive* loads. The development of complex GaAs MMICs will soon permit the implementation of active frequency converters with high dynamic range, selectable image characteristics, and conversion "gain" instead of an insertion loss. We should see these devices — up to 4-5 GHz — readily available in the not-too-distant future.

Plain old resistors and capacitors are taking on a new look. Chip- and surface-mounted components with superb characteristics well into the microwave region are now available. The fact that most modern circuits operate at low voltage and power levels means that spacings and insulation can be very small. This in turn reduces the overall size of the component, and hence its parasitic inductance and capacitance. This trend will continue to provide more components with nearly perfect RF characteristics in the VHF/UHF and low microwave ranges.

Even coaxial cables continue to undergo substantial improvement. Semi-rigid coax with foil shielding, foam dielectric, and non-contaminating jacket is available at very affordable prices. This type of cable is usable to 500 MHz with losses of less than 0.1 dB/meter. Small diameter hardline is now available to the UHF/Microwave enthusiast on the surplus market, thanks to the huge quantities used by military and cable TV applications. These cables are nearly the ultimate in transmission lines; they boast very low losses well into the microwave region, 100 percent shielding, and closed-cell foam insulations with permanent water intrusion barriers. Properly installed, these cables can have a useful life of more than 20 years.

Some of the least heralded advances in component design have taken place in the realm of interstage filters. Multipole ceramic and crystal filters with flat passbands, steep skirts, and low spurious responses have become available at remarkably low prices. Made for virtually all of the standard IF frequencies (455 kHz, 10.7, 21.4, 30, and 45 MHz), these marvelous filters, buried deep in our equipment, give us selectivity and freedom from unwanted responses that would have been a dream just a few years ago.

The advances continue. Monolithic crystal filters with narrow bandwidths (5-15 kHz) have been fabricated at over 250 MHz. Surface Acoustic Wave (SAW) filters with good characteristics up to 1.5 GHz are now available commercially. The availability of these higher frequency filters means that next-generation equipment can use up-conversion (IF higher than the incoming signal) to give us image-free, single conversion (large dynamic range) performance.

As the spectrum becomes more intensely populated, we are fortunate that component designers and manufacturers are investing in the technology to assure that there will be room for all of us.

ham radio

DECEMBER 1985 / $2.50

DIGITAL COMMUNICATIONS

Season's Greetings from . . .

ham radio magazine

hr

focus on communications technology

spread spectrum • a packet radio primer • AMTOR, AX.25, and HERMES • a packet radio frequency and deviation tester • W1JR, K4IPV, W6SAI, K0RYW • plus ham radio's five-year cumulative index

ham radio
magazine

DECEMBER 1985

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contents

W6MGI's column, "The Guerri Report," will return next month.

REFLECTIONS

another revolution

We are in the midst of another revolution, albeit a peaceful one. Packet Radio and other forms of digital communications have arrived. Previously staunch users of the various forms of AM and FM have found themselves drawn into the world of zeros and ones and are talking to each other with ever-increasing speeds, efficiencies, and applications. For example, on 220 MHz and higher, up to 5600 characters per second transmission link speeds are both theoretically and legally possible. Passing the same amount of data on CW would require a sending speed of approximately 60,000 wpm — almost a thousand times faster than Ted McElroy's long-standing Morse code record.

Of course the proliferation of personal computers has had an important influence on this process, but I still attribute the rapid growth of this field to the inquisitive, intelligent, and practical mind of the Radio Amateur. For as soon as the first generation of Terminal Node Controllers (the interface between the terminal and the radio) became available, user groups formed and began developing applications ranging from direct message-passing to Packet Bulletin Board Systems (PBBSs).

Though most Packet activity occurs on VHF (in large measure on 145.010 MHz), this does not restrict transmissions to short distances. In fact, as of this date, a large section of each coast is interlinked by a series of digital repeaters, or as they're better known, "digipeaters."

The startling growth rate of this mode became quite obvious as we prepared David McLanahan's article, "A Packet Radio Primer," for publication. Space was allocated for the April 10 version of the East Coast packet map shown on page 33. By September 10, activity increased dramatically and the size of the map *doubled*. This represents an increase of approximately 100 percent in the number of digipeaters, PBBSs, and home stations used predominantly for digipeating.

It seems particularly appropriate to consider this subject this month. December is, after all, a time of hope and renewal. In the past we've seen exciting developments — such as SSB, FM, and computers — make their mark on Amateur Radio. Will digital communications be the next logical step in this evolutionary process?

Turn the pages of this month's *ham radio* and see when and why the different forms of Packet Radio are appropriately used. And while you're on the subject, see how and why some hams have been experimenting with spread spectrum transmissions, once the exclusive domain of the military.

To paraphrase a well known soft drink manufacturer's slogan, "We're the Packet generation." Read on and see where you might fit in.

Rich Rosen, K2RR
Editor-in-Chief

presstop de W9JUV

THE AVERAGE U.S. AMATEUR IS JUST OVER 46 YEARS OLD, FCC's analysis of last April's Form 610s indicates. A detailed study of all that month's 9632 applications (all new licenses, renewals, and modifications) shows Novices to be the youngest, with an average age of 38.5. Techs average 45.1; Extras, 47.3; Generals, 50; Advanced, 51.8. By call areas, 9th district Novices were the youngest, at 35.6, and 0 district Advanced the oldest, at 55.

The U.S. Amateur Population Increased 0.65% During FCC's Fiscal 1985 (ending October 1), though the number of new Amateurs actually decreased by 7.6%. 17,373 newcomers joined Amateur ranks last year, while 14,709 dropouts left the year's end total 412,587. Largest percentage increase by license class was in Extras, up 2,344 to almost 38,000.

SOME MEANS SHOULD BE ESTABLISHED FOR "CERTIFYING" FREQUENCY COORDINATORS on a state or regional basis, as there's apparently no interest in establishing a "National Coordinator," FCC Safety and Special Services Chief Ray Kowalski suggested in his comments during the FM forum at the ARRL National Convention in Louisville. "Certified" coordinators would be those established and generally recognized for a given area; an eight-point plan is to be developed for determining appropriate qualifications and the method of formalizing such certification. Though the plan is to be published in the ARRL's Repeater Coordinators' Newsletter when completed, the League is specifically NOT participating in developing the plan.

Texas's Recoordination To 20 kHz Spacing On 2 Meters Has Been Completed, with the actual shift throughout the state expected to be complete by the end of November.

ARRL'S PROPOSAL TO PERMIT NOVICES PHONE PRIVILEGES probably won't be worked on at the FCC until early 1986, meaning that an NPRM won't be out until late spring at the earliest. So far there seems to be considerable division in the Amateur ranks on the issue, but the age figures in our lead item certainly indicate some change is needed.

Japanese CBers Have Invaded The Lower End Of 10 Meters, P29JS reports in a letter to the Southern California DX Bulletin. He's been hearing Japanese chatter all the way up to 28585 kHz, but without any indication of any "JA" callsigns.

METROPLEX HAS BEEN NAMED THE FOURTH NATIONAL VEC effective September 19, the FCC has announced. Metroplex, one of the very first regional VECs, is now actively seeking VEs in other call areas. Call Alex Magocsi, WB2MGB, at (201) 592-6243 for information.

18 Regional And Three National VECs Have Expressed Interest In Joining CARE (Council of Amateur Radio Examining), which is now well along the road toward becoming a viable organization and expects to be incorporated as a not-for-profit corporation under Illinois law before the end of the year. Jim Georgias, W9JUG, can provide further information.

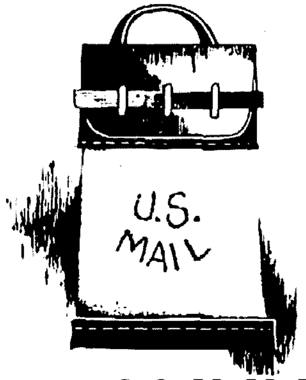
AFTERSHOCKS FROM NEWS MEDIA EXPLOITATION OF AMATEUR FREQUENCIES during the Mexico City earthquake disaster are still going on, with both the FCC and some of the offending news organizations on the receiving end of complaints by concerned Amateurs. Even some media people have themselves admitted feeling that the situation went far beyond reason, though most did not want to be quoted on the issue. Preliminary discussions between the FCC and both the Radio and TV News Directors Association and the National Association of Broadcasters have reportedly taken place, with hope that acceptable guidelines for Amateur Radio/media cooperation can be set up before the next crisis occurs. The FCC had thought it had defined the limitations in its Report and Order on BC Docket 79-47, but a widely distributed industry interpretation of the FCC's action left many in the industry with the impression that Amateur Radio was for their use pretty much as desired!

The FCC Is Interested In Reports Of Specific Media Incursions during the Mexico City crisis; they must be first-person and specific enough to be related to a specific news organization. Send them to Raymond Kowalski, Chief, Safety and Special Services, FCC, 1919 M St., NE, Washington D.C. 20554; tapes of abuses would be particularly welcome.

The Issue Of "Non-Amateur" Use Of Amateur Frequencies is an on-going issue that can't be ignored; the recent petition by a low-power TV broadcaster to use frequencies on the 70-cm Amateur band for TV remotes, and the establishment of a "protection zone" along the Canadian border for Canadian 420-430 MHz commercial users, are cases in point.

STILL MORE "HAM IN SPACE" OPERATIONS FROM SPACE SHUTTLES are shaping up for next year. AMSAT member Dr. Ron Parise, WA4SIR, has been selected to fly on Mission 61E in March, while Dr. Owen Garriott, W5LFL, is scheduled for mission 61K, now set for next September.

GEOSYNCHRONOUS AMATEUR SATELLITES ARE STARTING TO LOOK like real possibilities, according to AMSAT. Earlier, NASA had said it might provide Amateur Radio capability on one of its Advanced Communications Technology Satellites (ACTS), and now it appears an AMSAT transponder ("Phase 4") may find a spot on the same bird. In addition, Arianespace may also be able to provide a piggy-back launch opportunity for an Amateur transponder into geosynchronous orbit. AMSAT officials are now reviewing the possibilities, with particular interest in incorporating new approaches and capabilties in an Amateur proposal.



comments

Dear HR:

In response to KAØDOE's letter in the September, 1985, issue of *ham radio* (Comments, page 15) I have only one comment. The total function of the Novice Amateur Radio license is to introduce the public to Amateur Radio Service by giving them simple privileges on a few selected portions of the Amateur Radio bands and a very simple but effective mode of transmission called *CW* or Radiotelegraphy. Then it's up to the individual to make that committment to upgrade to the higher grade license and more privileges.

Bill Eaton, WB1CXI
Channelview, Texas

something for nothing

Dear HR:

Unlike KAØDOE, I read July's "Presstop" (page 6) with horror and disgust. The present-day attitude of rewarding hams for doing absolutely nothing will be the biggest reason for the ultimate death of the hobby.

Why are unearned extensions of band privileges the only thing that can help encourage these operators? Aren't there myriad challenges available to try the patience and ingenuity of any class of Amateur today? Will more "entertaining" modes give us any "more to encourage us to seek out our full potential as operators?"

At the beginning of World War II, Amateur Radio operators provided a highly proficient cadre of electronics technicians and skilled instructors to the United States. Let me propose a calamity today. What functions could a group of Amateurs provide whose only claim to technical skill is the ability to box up their equipment and get it to the United Parcel Service?

I was granted a Class B license in 1947 and, one year later, passed the Class A examination. Can I now cry out that I have been a good and faithful ham for 38 years, so I deserve an Extra Class license? Hell, no! I must get my old brain in gear and hit the books and suffer my way back to 20 words per minute. That is one of the basic challenges of Amateur Radio. The frightening thing about "glorified CB operators" is not only the terrible operating practices that a lot of them demonstrate, but the "something for nothing" syndrome that fosters these practices.

If a soldier received the Purple Heart for falling off the back of a 6x6 when he was drunk, it would degrade the sacrifice of all the others who proudly wear that decoration. The time and effort spent to upgrade a license is vital to the spirit of Amateur Radio. With it comes the understanding of the privilege granted us to use certain parts of the radio spectrum. And with that understanding comes the resolve to operate legally and properly.

I raised my children to realize that they were due only what they were willing to expend their time and effort to gain. There is, and certainly should be, no free lunch. In like manner, I do not condone any unearned privileges for any Amateur Radio operator.

Joe Weite, KH6GDR
Makakiio, Hawaii

short circuit

75-meter transceiver

The PC board art shown in K1BQT's article, "A Compact 75-meter Monoband Transceiver" (November, 1985, page 13) is incorrectly sized. For a complete, corrected set of board art suitable for reproduction, send a stamped, self-addressed No. 10 (business-sized) envelope to *ham radio*, Greenville, New Hampshire 03048.

spread spectrum and digital communication techniques: a primer

Not quite sure how speech can be transformed into a digital signal? *Read on.*

Ever since the inception of Amateur Radio, hams have kept abreast of the latest and most innovative methods of communication. From the advent of spark-gap radiotelegraphy, to the early FM transmitters, through the 2nd World War — when hams played an important part in concocting the first reliable pulse radar systems — to RTTY, SSTV, SSB, satellite communications, and packet radio, Amateurs have been ardent users of new and fascinating modulation methods.

In recent years there has been interest in a relatively new type of communication technique. While the foundation for this communication method was laid with the advent of ranging radar, it has only been in the past 10 to 15 years that it has received so much attention from both the military and the private sector. This technique, known as *spread spectrum* (SS) is unlike any communication method previously tried by Amateurs. However, judging by our track record, it would seem that it's only a matter of time before we familiarize ourselves with it.

The purpose of this article is to provide an overview of spread spectrum communications for those not familiar with it. While the topic is much too broad to be fully discussed here, the major concepts will be highlighted in a manner that can, I hope, be understood by those with little math background. As well as the concepts governing spread spectrum and digital communications, typical station hardware requirements will also be addressed.

why spread spectrum?

One might wonder how and why spread spectrum evolved, and how it could be applicable to the Amateur Radio bands. To address the first query, it is necessary to consider the problems associated with military communications during World War II. At the time, jamming and antijamming techniques were the order of the day. By 1945 every Allied Bomber plane was equipped with two jamming transmitters, while it is estimated that as many as 90 percent of all electrical engineers in Germany were involved in an antijamming program of gargantuan proportion.[1]

To combat the effects of jamming, spread spectrum was used to spread the signal out, thereby rendering narrow band jammers virtually ineffective. In addition, the fact that spread spectrum could be used with a *low probability of intercept* (LPI) made this an ideal method of communicating while appearing "radio silent" to conventional receivers.

Today the quest for a signal that cannot be jammed continues in military circles; commercial applications, such as banking and private mail systems, require security. As jamming and intercept capabilities become more sophisticated, methods of communicating become increasingly complex. Spread spectrum continues to evolve into a highly complicated mode of communication. Those fascinated with the history of radio would find the accounts of the development of spread spectrum to be very exciting reading. Several excellent accounts are listed in the references.[1,2,3]

The above-mentioned attributes hardly seem appropriate for Amateur Radio! The FCC rules prohibit any kind of coded or secure communications, and intentional jamming is a problem we would ideally never

By Ted S. Rappaport, N9NB, Box 283, Electrical Engineering, Purdue University, West Lafayette, Indiana 47907

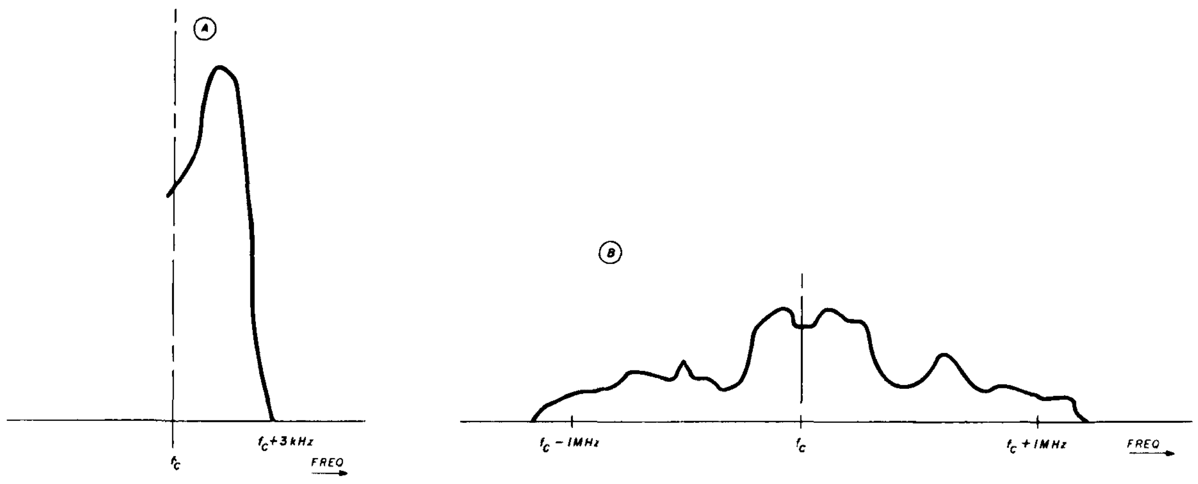


fig. 1. Comparison of bandwidths between narrow modulation (*A*) and spread spectrum (*B*).

have to deal with. There are several other benefits, however, that might be of use to us in the future as the Amateur Radio spectrum becomes saturated with users. In fact, a look at why mobile telephone companies are considering SS sheds some light on some of the possible rewards.

In metropolitan areas, where there are many mobile telephone users in a small area, cellular radio has been introduced to alleviate the congestion in the mobile telephone spectrum. In a cellular radio system, as the term suggests, the city is broken into "cells," with each cell having its own multichannel repeater capable of handling a limited number of users within the cell.[4] As the user travels into an adjacent cell, the adjacent repeater takes over the communication. The cellular technique has been used to increase the maximum number of mobile telephone and commercial radio users from several hundred to several thousand in many cities across the country. Of special interest to the industry is the fact that compared to conventional narrow band modulation techniques, SS has the capability of supporting a larger number of users for a given cell size![5]

Other advantages that SS offers to both the military and the mobile communication industry include selective addressing capability, code division multiplexing, and interference and multi-path rejection. With SS, it is possible for a transmitter to selectively communicate to one or several receivers while remaining oblivious to other users. Also, several stations may use the same band of frequencies simultaneously without interfering with one another. Since SS signals have very wide bandwidths, conventional narrow band users may also use the same spectrum without adversely affecting the SS communication. Conversely, the average power of a SS signal in any narrow band region is small, so the narrow band modulation is not severely QRM'ed, either.

As will be demonstrated shortly, SS requires more complex hardware than does conventional narrow band equipment; as the spectrum stands today, its use in Amateur Radio is probably not currently warranted. However, with increased HF and VHF/UHF activity, it is conceivable that we may eventually need a drastically different approach to communications. Progress has recently been made by hams in such areas as coherent CW and packet communications.[6,7,8,9] The inevitable thrust toward digital communications makes SS appear to be a likely modulation method in the future.

overview of spread spectrum

Figure 1 illustrates the bandwidth of an SSB speech signal compared to a typical spread spectrum signal. As its name implies, spread spectrum is a modulation method whereby the energy of the transmitted signal is spread out over a very wide bandwidth. This is quite unlike SSB or narrow band FM (NBFM), where the transmitter output has a bandwidth on the order of that of the modulating signal (the usable audio bandwidth for speech is about 3 kHz). However, wide band FM (WBFM) transmitters have bandwidths that are many times greater than that of the modulation. Clearly, though, WBFM is not spread spectrum! This is where the second important distinction between spread spectrum and conventional modulation methods must be made.

A spread spectrum communication system uses a

special generated wide band signal that is independent of the message modulation. At the transmitter end, the message modulation (a voice signal) may be multiplied by this independently generated signal, and the resulting mix then transmitted on a carrier. This is known as *Direct Sequence Spread Spectrum* (DS). Another type of SS, known as *Frequency Hopping Spread Spectrum* (FH), can be generated by using the independently generated signal to cause the carrier signal to frequency hop in a prescribed manner.

It is the independent signal that determines the amount of bandwidth spreading at the transmitter output. It also determines the immunity the SS signal has to narrow band interference. This independent signal is always digital, and is generated by digital logic devices (such as TTL). The term *pseudo noise code* (PN) is used to describe this independent signal since to an uninformed observer the PN code looks like a random jumble of 1s and 0s. Actually, though, the PN code is a periodic sequence that can be easily gener-

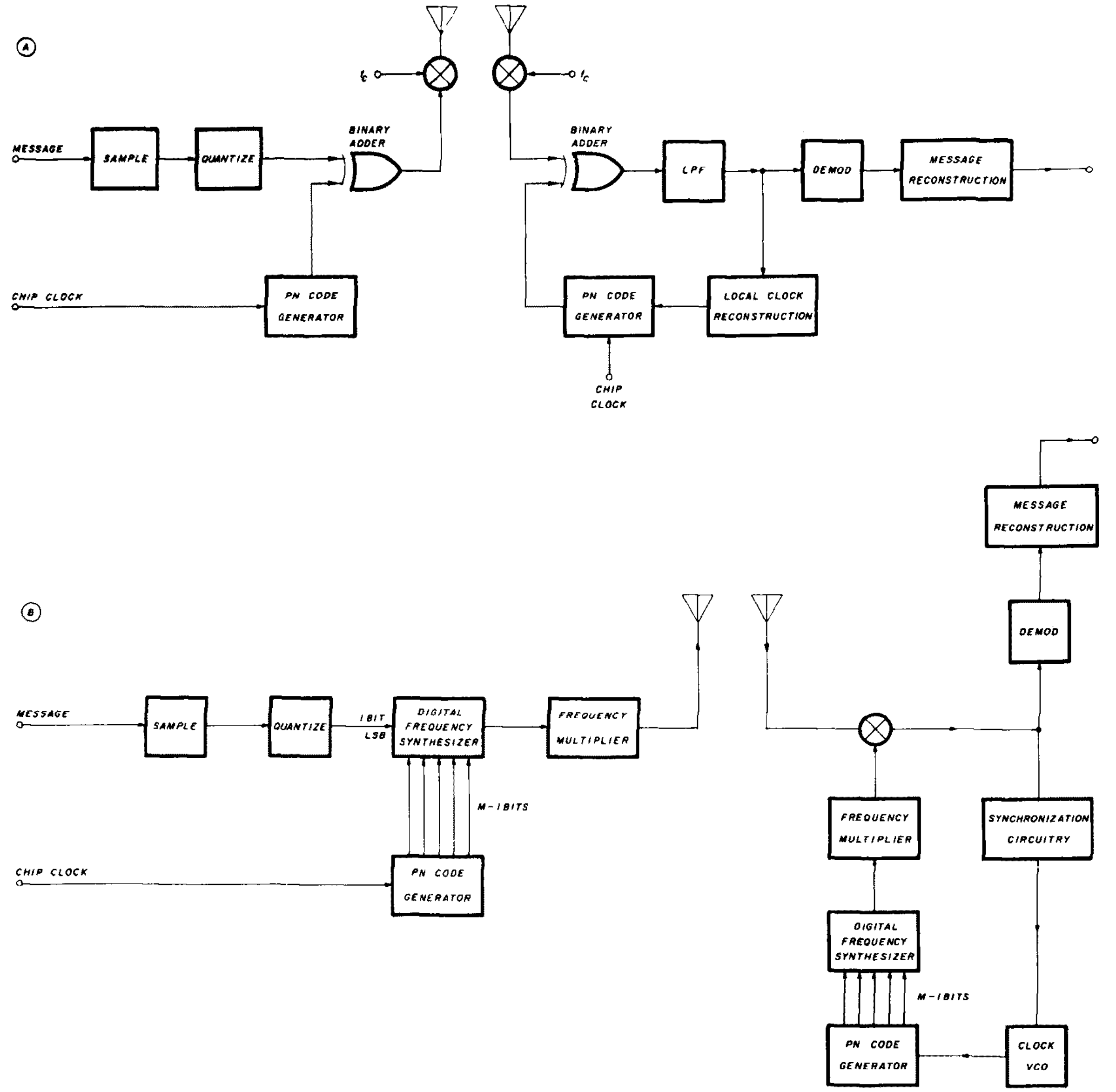


fig. 2. Spread spectrum transmitter/receiver block diagrams: (*A*) direct sequence (DS) transmitter and receiver, (*B*) frequency hopping (FH) transmitter and receiver.

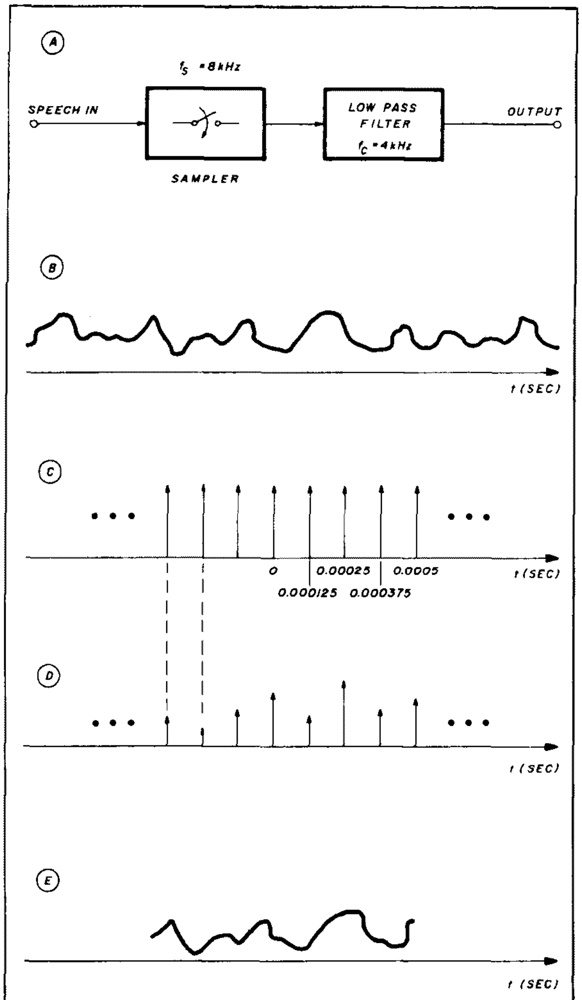


fig. 3. Digital sampling-time domain analysis: (*A*) digital sampling system, (*B*) speech signal (f_{max} = 4 kHz), (*C*) representation of sampler action, (*D*) output of sampler, and (*E*) output of low-pass filter (reconstructed message).

ated by a sequence of shift registers. In order to recover the original message, the receiver must be able to reconstruct the same PN code used by the transmitter. When the transmitter and receiving encoding signals are identical, and when they are synchronized in time, then the message is detected. **Figure 2** illustrates block diagrams of spread spectrum transmitter and receiver pairs for both the DS and FH case. A more detailed look at the generation of the PN code is considered subsequently. However, before delving into the details of SS, it is first necessary to become familiar with some basic concepts of digital communications.

digital communicaton concepts

Because the encoding signal is digital, spread spectrum can be considered to be a special form of digital communications. Unlike SSB, AM and FM, which are analog, continuous time communication methods, digital communication systems work on the principle of the sampling theorem.

The sampling theorem, developed by Nyquist in 1924,[10] states that a continuous time signal may be represented by a sequence of discrete time snapshots, or samples, without any loss of information in the signal, provided that the samples are taken at a rate which is at least twice as great as the highest frequency component of the original continuous time signal. A basic relationship which relates the sampling frequency (f_S) to the time duration between successive samples (T_S) is

$$f_s = \frac{1}{T_s} \tag{1}$$

Figure 3A shows the components of a typical sampling system. **Figure 3B** illustrates a typical speech signal that has been band limited to have a peak frequency component of 4 kHz. The action of the sampler is shown in **fig. 3C**. **Figure 3D** illustrates the output of a sampler that is taking samples at a rate of 8000 samples per second (twice the rate of the highest message component). **Figure 3E** shows the recreated message waveform after the samples are placed through a low-pass filter having a cutoff of 4 kHz.

In order to lay a foundation for the analysis and understanding of SS, it is instructive to look at the sampling theorem from a different point of view. In the early 1800's, Fourier, a famous mathematician, observed that most functions could be represented by a summation of sinusoids having different amplitudes and periods. In short, he laid the groundwork for the development of the celebrated Fourier transform. This transformation allows one to analyze a signal in the frequency domain rather than in the time domain.

Frequency domain analysis can directly give information pertaining to the bandwidth of a signal. Tables such as **table 1** have been compiled which lists the Fourier transforms of many common signal shapes.[11,12,13] By analyzing the sampling circuit of **fig. 3A** in the frequency domain, we can better explain how and why the sampling theorem holds.

From **table 1**, the Fourier transform of an impulse sampler is an impulse train in the frequency domain. Note that in the time domain (**fig. 3C**), the sampling action is effectively multiplying the input signal by a "1" at each sampling instant, and multiplying by "0" in the interval between samples. Just as time signals have Fourier transforms, so do time operations such as addition and multiplication. The Fourier transform of a time multiplication is known as *frequency convolution*. Convolution is a fundamental concept in control

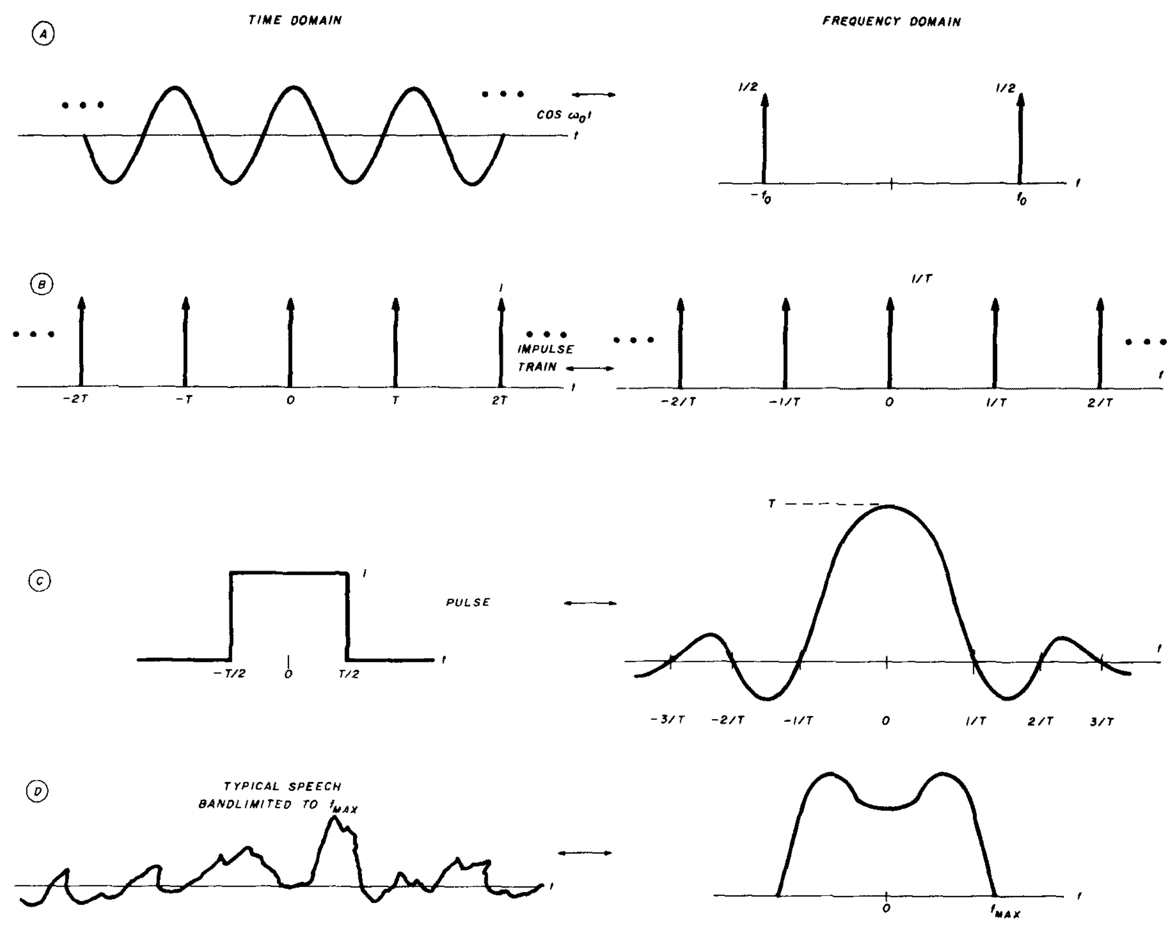


table 1. Time functions and their corresponding Fourier transforms.

and communication theory, and is used to express the output of a system or filter in terms of the input signal and the system impulse response, (i.e., the response of the system to a sudden input signal). For this discussion, it is necessary to know only that the convolution of a band-limited spectrum (**fig. 4A**) with a frequency pulse train (**fig. 4B**) yields the original message spectrum replicated at each of the pulse train harmonics. Hence, by frequency domain techniques, we find that the output of an ideal sampler is the original message spectrum replicated throughout the entire frequency domain and separated by integer multiples of the sampling frequency (**fig. 4C**). By low-pass filtering the sampler output, we can recreate the original message exactly (**fig. 4D**). By the same token, we could bandpass filter the output of the sampler and also recreate the message, although this is not usually done.

If the highest frequency of the input message exceeds one half of the the sampling rate, then an undesirable effect known as *aliasing* occurs. As can be seen in **fig. 4E**, each adjacent message spectrum overlaps so that the LPF output is not the original message, but rather a distorted signal. With frequency domain analysis, it becomes clear why the sampling frequency must be at least twice that of the peak frequency component of the input.

Before moving on, it should be noted that this quick look at the sampling theorem assumes an "ideal" sampler — one for which the sample durations are infinitely small. In reality, the sample durations are small, but finite. Taking this into account yields similar, but slightly more complicated, results. **Figure 5** illustrates the spectrum of the output of a typical "real world" sampler. As the sample widths become wider, there is less energy at the higher frequencies. This is why the sampler is followed by a low-pass filter rather than a band pass filter. Also neglected here are some amplitude scaling factors that are involved in transforming between the time domain and frequency domain.

These subtleties *are* required in exact problem solving, but are not important in gaining a good understanding of the sampling theorem. Those interested in the finer details of the sampling theorem and Fourier transform techniques might find the references helpful.[11,12,13,14,15]

data communication

Certain digital communication systems such as RTTY and packet radio, where the message text is originated by a keyboard rather than continuous-time speech, are known as data communication. In this case the sampling theorem does not apply, since there is no continuous time signal to sample. However, the data rate (the rate at which information can be sent) is a function of the number of bits used to represent each character, and is also a function of the time duration of each bit.

For example, the American Standard Code for Information Interchange (ASCII) prescribes that each keyboard character be represented by a unique 7-bit data word. The letter i, for example, is represented by the binary word 1101001. If each bit has a time duration of 1 millisecond, then any character may be sent down a channel in a time of 7ms. If a start bit, a stop bit, and a parity bit are sent along with the data, then one character can be sent every 10 ms. The data rate for this set-up would be 1000 bits per second (bps), or 100 characters per second.* Data communications of this type are termed "asynchronous" since the receiver never knows when the sender will depress the keyboard. The start bit and stop bit are the necessary overhead to identify each of the keyboard entries. The parity bit is used to validate the received data. **Figure 6** illustrates a complete asynchronous character word for the letter i.

There is a trade-off between bit rate and occupied bandwidth of a digital signal. A good rough estimate is that the required bandwidth of a digital signal is equal to the reciprocal of the bit duration. For the previous example, the required bandwidth would be on the order of

$$BW = \frac{1}{t_{bit}} \quad \frac{1}{0.001\ s} = 1000\ Hz$$

For a faster data rate, more bandwidth is required. In a band-limited channel, such as a commercial telephone line, there is an upper limit on the bit rate. This is why home computer modems seldom exceed a data rate of 1200 bps.[16] While spread spectrum is well suited for either voice or data messages, the remainder of this article will consider only a voice message. Once the voice is "digitized," it is sent through the channel in the same manner as data.

*In digital communications the bit rate is the same as the baud rate, so for this example the data rate is equivalently 1000 baud.

quantization

From the sampling theorem, we know that it is necessary to send only the voltage values of the samples rather than the continuous time signal. If the sample values could be represented in a digital fashion, then we could take advantage of schemes that have been developed expressly for digital communication. In short, digital communications systems are able to outperform analog methods because signal reception is based on distinguishing whether a "0" or a "1" was sent, rather than trying to recreate a random continuous time waveform directly. Schemes such as *error correction coding* and *minimum probability of error receivers* can be used to provide far superior performance when compared to analog communication techniques.

To represent the height of a sample value digitally, it is necessary to quantize the sample. For binary data (standard digital logic), the quantizing action truncates the actual sample value so that each sample is represented by a fixed number of bits (i.e., every sample is expressed by N bits) in the time between successive samples. Since it is conceivable that the samples can take on a continuum of values, there is some error introduced by the quantizer. However, if a limiter is used (to contain the voice signal voltage within the limits of the quantizer) and if there is a sufficient number of levels in the quantizer, then this error, known as *quantization noise*, is quite small.

The quantizer is an important concept in digital communications. The resolution, or the accuracy in which a sample can be represented, is directly related to the N, the number of bits used to represent each sample. For N bits, there are 2^N quantizer levels (sometimes called *bins*).

An example is useful to clarify how speech can be quantized and sent down a channel as a digital bit stream. **Figure 7** demonstrates how the sample values are assigned data words in a three-bit quantizer (N = 3). As indicated in **fig. 7**, there is some error introduced because of the fact that each sample is represented by only a three-bit word. By using more bits, each sample can be more accurately represented. Surprisingly, though, even with only three bits of quantization, intelligible speech can be transmitted.[17] Once assigned a quantization data word, the truncated sample is converted into 1s and 0s and sent down the channel. The output of the quantizer is known as *pulse code modulation*, since the message has been coded into a train of digital pulses. Since N bits must be sent in the time between adjacent samples, the bit duration of the quantized data is

$$T_{bit} = \frac{T_s}{N} \qquad \textbf{(2A)}$$

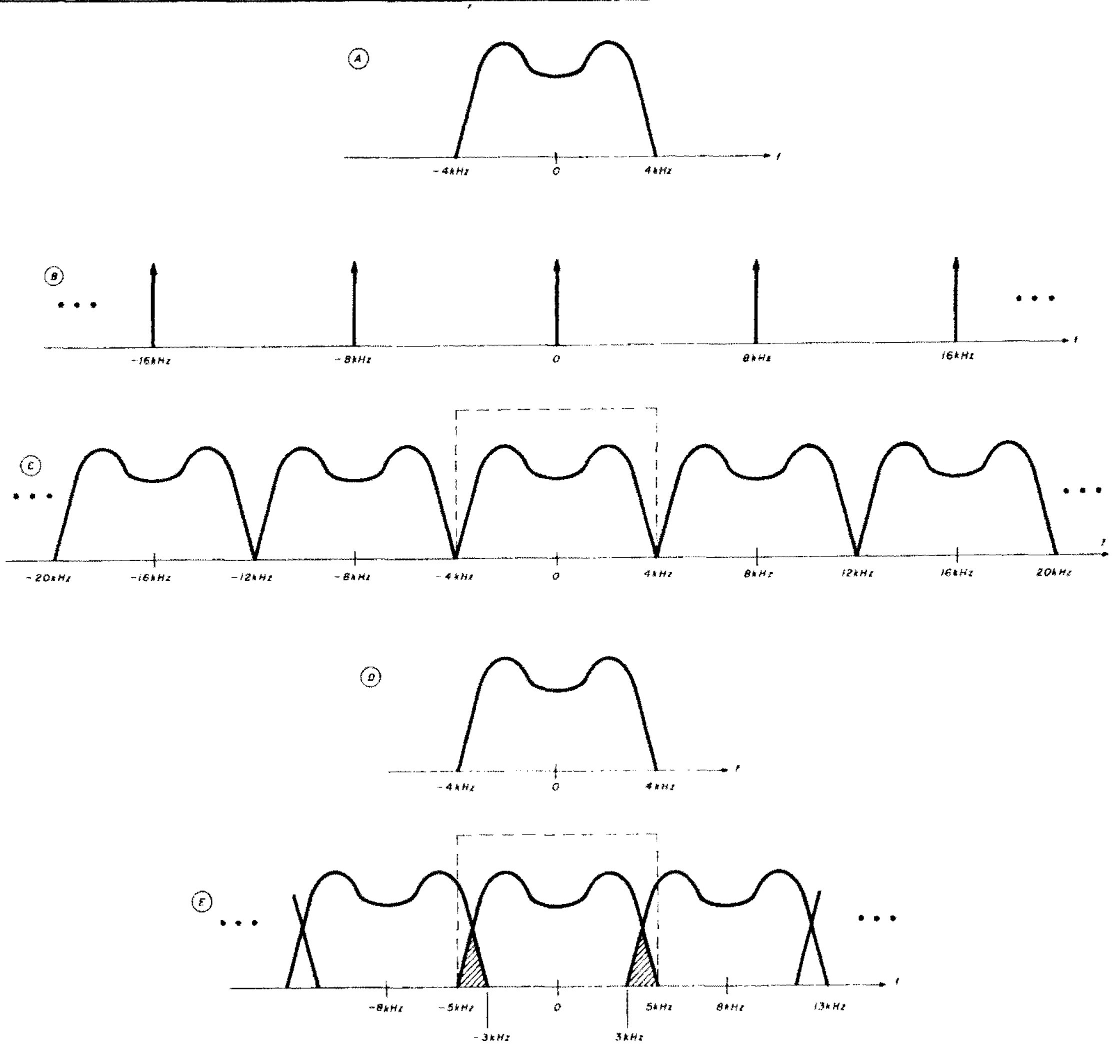


fig. 4. Digital sampling — frequency domain analysis: (*A*) spectrum of speech signal (f_{max} = 4 kHz), (*B*) spectrum of sampler action, (*C*) output of sampler, (*D*) output of recreated message spectrum, and (*E*) aliasing example, f_s = 8 kHz and f_{max} = 5 kHz.

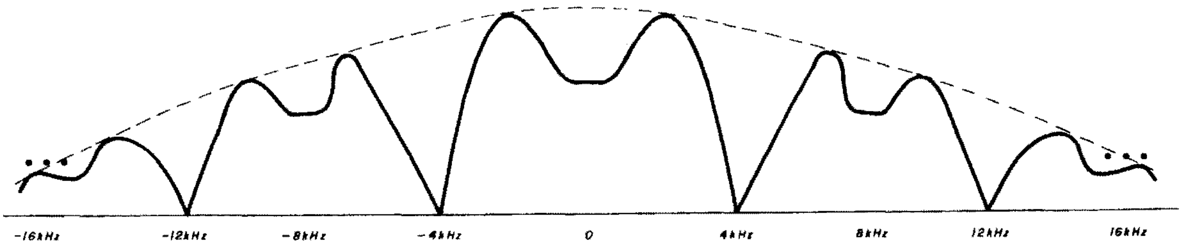


fig. 5. Spectrum of output for non-ideal sampling.

where T_S is the time between adjacent samples. Since

$$T_S = \frac{1}{f_S} \quad \textbf{(2B)}$$

each quantizer bit duration is given by

$$T_{bit} = \frac{1}{Nf_S} \quad \textbf{(2C)}$$

Hence, the bandwidth is given by

$$BW = Nf_S \quad \textbf{(2D)}$$

A practical method of sampling and quantizing is to use a *sample and hold* circuit followed by an *analog-to-digital* (A/D) converter. The sample and hold is similar to the ideal sampler, except the sample height is held for the entire time duration between samples, and is updated at each new sampling instant. While the sample value is held at a constant level, it is converted into a digital signal by the A/D converter. The end result is identical to that of **fig. 7**.

To recreate the message, the digital bit stream is clocked into a *digital-to-analog* (D/A) converter. This undoes the effect of quantizing which the A/D had

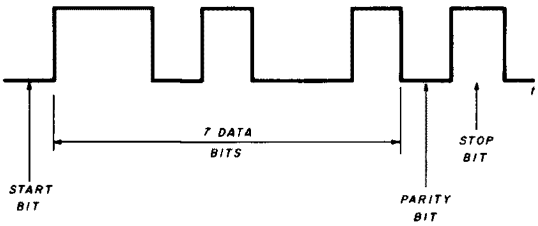


fig. 6. ASCII 10-bit asynchronous data character for the letter i.

upon the original message samples. The D/A output is then low-pass filtered to transform the reconstructed samples into the original message.

spread spectrum systems

There are many types of spread spectrum systems. These include direct sequence (DS), frequency hopping (FH), time hopping (TH), chirp, and hybrid systems which combine several techniques at once. Only DS and FH are considered here, since these seem to be most easily implemented. Common to both of these types of SS systems is the need to generate and reconstruct a PN (pseudonoise) code.

To produce the DS spread spectrum signal, a PN code signal must be produced that has a bit rate (bandwidth) much greater than that of the quantized message. For FH spread spectrum, it is not so much the bit rate that matters as does the number of bits used in a complete cycle of the PN code.

The PN code can be generated by a feedback arrangement of flip-flop stages. A flip-flop is a digital device which can store a binary value (either a 0 or a 1). Flip-flops can be connected in series to form shift registers. As the term implies, shift registers store several binary digits and shift them to the left or right each time an external clock pulse is received.

The simplest PN codes (there are several types) are known as *maximal linear codes*, or *m-sequences*. These are produced by m-stage shift registers which use feedback to produce periodic codes that have N bits before recycling. For an m-stage shift register, there are $N = 2^m - 1$ bits in each period. **Figure 8** illustrates a three-stage shift register. All three flip-flops of the shift register are clocked to the right simultaneously, and each time a clock pulse appears, a new binary digit appears at the output. **Table 2** indicates the value held by each flip-flop for a given time inter-

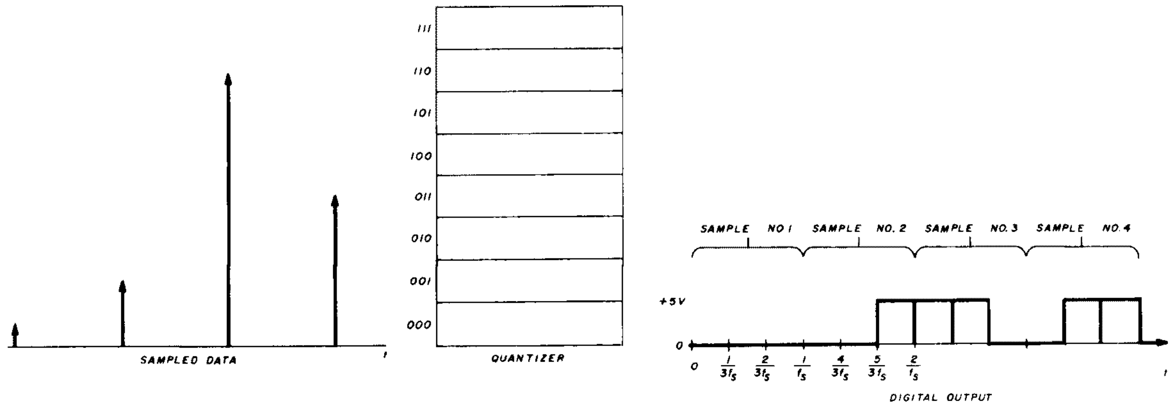


fig. 7. A three-bit word used to quantize sampled data.

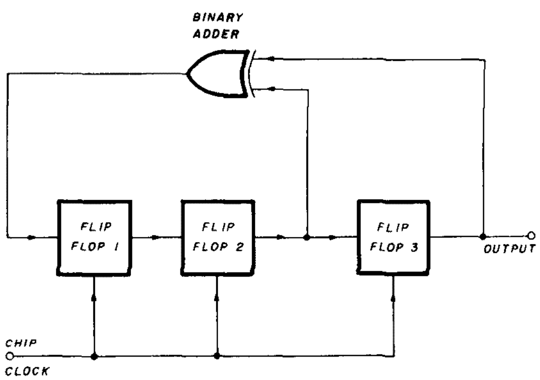


fig. 8. Shift register generator for M=3.

table 2. Contents of Shift Register Stages for m = 3.

	stage 1	stage 2	stage 3
initial contents	1	1	1
clock pulse 1	0	1	1
clock pulse 2	0	0	1
clock pulse 3	1	0	0
clock pulse 4	0	1	0
clock pulse 5	1	0	1
clock pulse 6	1	1	0
clock pulse 7	1	1	1

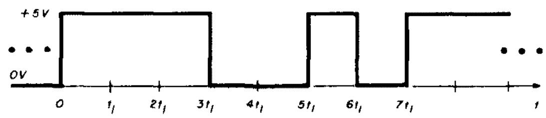


fig. 9. Digital output of three-stage shift register.

val. **Figure 9** shows the digital signal that would be produced by a standard TTL circuit.

To avoid confusion between the PN code and the actual message bit stream, the term chip is used to describe each bit of a PN code. For a chip duration of t_1 seconds, the periodic PN sequence repeats itself every Nt_1 seconds. For the example shown in **fig. 9**, the same chip value would be seen every $7t_1$ seconds apart.

By changing shift register feedback paths, it is possible to generate many unique m-sequence codes, each having a period of N chips. The number of possible codes is important since it defines the maximum number of users that can be uniquely addressed, assuming that each user has the same length shift register. The exact number of unique codes is dependent upon the number of shift register stages and the possible feedback paths. In general, the longer the shift register, the more unique codes exist. However, as can be noted in **table 3**, when the shift registers consist of a prime number of stages, there is a maximum of codes for a minimum of hardware.

Modular Shift Register Generators (MSRGs) such as the MC8504 are available for easy PN code generation. The MC8504 is a 16-pin chip that features four stages of an expandable shift register. They may be cascaded, and additional flip-flops (such as a 7474) may be added to implement an arbitrarily long m-sequence. A nine-stage MSRG capable of producing 48 selectable 511 chip codes is shown in **fig. 10**.

Direct sequence (DS) spread spectrum. As shown in **fig. 2A**, the PN code is added with the digitized message (usually PCM) to produce a digital signal that can be readily modulated. If one of the adding signals is wide band, then the resulting adder output signal is also wide band. Since the PCM has a bandwidth on the order of the original message, it is necessary to use a PN code which has a bandwidth several orders of magnitude larger than the message bandwidth in order to obtain large bandwidth spreading. For this reason, the chip rate is typically run at speeds of several Megachips per second (Mcps).

The output of the binary adder is fed into a balanced modulator. This modulator produces a particular carrier phase for a logic "1" and a 180-degree shifted carrier for a logic "0." Hot-carrier diodes are used in conjunction with wide band transformers to produce the final RF signal. **Figure 11** illustrates a double-balanced mixer, which is the most commonly used type of balanced modulator. The balanced modulator input and output signals are shown in **fig. 12**. While the output waveform appears simple in the time domain, frequency domain analysis reveals that there is a wide band of frequency components centered around the nominal carrier frequency.

In a DS-SS receiver, just as with any RF receiver, it is first necessary to bring the modulation down to baseband. This is accomplished by mixing the received signal with a local oscillator that is adjusted to the transmitter's frequency. Then, the coded signal must be matched with an internally generated PN code. This is accomplished by binary addition. If the internal PN code generator is not correlated with the incoming signal, the resulting adder output is called *code noise*. However, when the internal code is synchronized with the incoming signal, the output of the binary adder collapses to the original digital message bit stream. A manual tuning dial can be used to adjust the phase of the code generator until synchronization is obtained. Better yet, a microprocessor can be used to automatically adjust the receiver PN code phase. Once synchronized, the demodulation may be accomplished by using a D/A converter followed by a low pass filter.

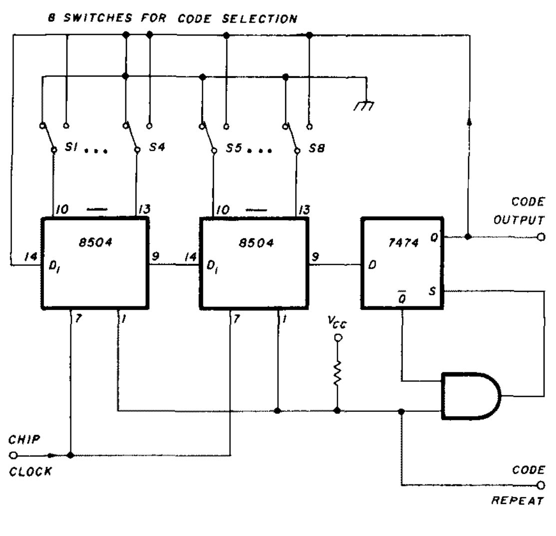


fig. 10. Implementation of nine-stage MSRG.

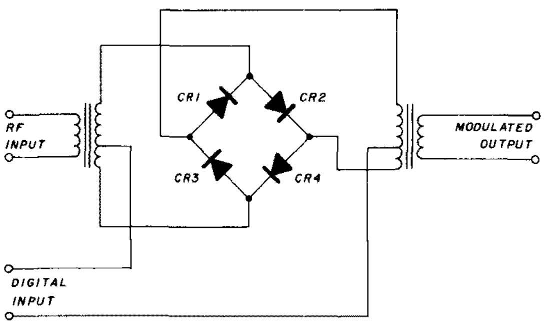


fig. 11. Double-balanced mixer.

table 3. Characteristics of M-sequences of various length.

M stages	maximum number of unique codes	number of chips per period
3	2	7
4	4	15
5	6	31
6	4	63
7	18	127
8	16	255
9	48	511
10	60	1023
11	176	2047
13	630	8191
20	19,200	1,048,575
30	11,880,000	1,073,741,823
61	3.1×10^{16}	2.3×10^{18}

Since the receiver uses an internal clock signal for code reconstruction, the low pass filter could be of the switched capacitor variety.

It is important to note several points of practical interest. Obviously, the receiver must tune to the sender's transmitting frequency in order to establish the possibility of communication. This suggests that calling frequencies would be advisable for the first Amateur attempts at DS-SS communication. Furthermore, a standard shift register length (or a few agreed upon lengths) seems mandatory, since the senders and receivers must be able to match each other's PN code. A fixed chip duration is necessary, too, so that all stations would be ensured that they could synchronize with the other users.

Before leaving DS-SS, a word should be said about PN code length. In military applications, where security is an important consideration, it is not uncommon to find PN coding schemes which use shift register stages of length 40 or greater. If each chip duration is 1 microsecond, then one cycle of an m = 40 PN code is completed in a time of

$$(2^{40} - 1) \text{ chips} \times \frac{10^{-6} \text{ sec}}{\text{chip}}$$
$$= 1.1 \times 10^{6} \text{ sec} = 12.7 \text{ days}$$

Even if an interceptor listened in on this signal for several hours, it would appear to be a random jumble of binary digits. For long codes such as this, it takes a very long time to synchronize the receiver. On the other hand, if a shorter PN code is used (say, m = 13), then the entire code sequence is repeated every 8.3 milliseconds! A code having such a short period can be synchronized quite quickly at the receiver end.

Frequency hopping (FH) spread spectrum. In a FH-SS system (**fig. 2B**) there is a narrow band transmission occurring at any given time instant. However, there is a wide range of frequencies from which the transmitter may select. The particular frequency selected for use at any given moment is determined jointly by the digitized message bit stream and the PN code generator.

The message bit stream is used as the least significant bit (LSB) of an M bit data word. The PN code generator supplies the other M-1 bits. The data word is then used to determine the additive offset required to generate the proper frequency. Frequency hopping takes place over $N = 2^M$ frequencies separated by integer multiples of f_1, where f_1 is the gap between adjacent hop frequencies. The repetition rate of the frequency hopping sequence is determined by the number of stages in the PN code generator and by the speed of the chip clock. Since the generator supplies M-1 chips for each hop, the frequency hop sequence repeats every

$$\frac{(2^{m \text{ stages}} - 1)}{(M-1)} \quad \text{hops}$$

table 4. Indirect frequency synthesizer ICs and their characteristics.

part number	number of frequencies	reference frequency	maximum divider input frequency	control
Hughes HCTR0347	45	50 Hz-500 kHz	10 MHz	8 bits parallel
Nitron 6410	100	4.00 MHz	1.6 MHz	8 bits parallel
Motorola 145104	256	10.24 MHz	4.0 MHz	8 bits parallel
National DS8906	16384	10.24 MHz	120 MHz	20 bits parallel
National MM55110	1024	10.24 MHz	3.0 MHz	10 bits parallel
Fairchild 11C84	128	10.24 MHz	20 MHz	7 bits parallel
AD-TECH FS-2574	1000	10.00 MHz	258 MHz	10 bits parallel

Figure 13 illustrates a typical FH transmission for the case of $M = 3$ and a nine-stage m-code generator using a 1-kHz chip clock.

Typical values for a suitable FH system might be $f_1 = 500$ kHz and $M = 3$ bits. For this example, the total RF bandwidth of the system would be

$$2^3 \textit{ frequencies} \times \frac{500 \textit{ kHz separation}}{\textit{frequency.}} = 4.0 \textit{ MHz}$$

If the lowest frequency of the transmitter were 420 MHz, the highest frequency used by this system would be 424.0 MHz.

The frequency synthesizer is the key to an FH-SS system. Its operating characteristics (such as frequency range, switching speed, and hop duration) determine a system's capability. There are two major classes of synthesizers, the direct type and the indirect type. The direct frequency synthesizer uses filters and mixers and is seldom found in current Amateur gear. The indirect type uses phase-lock-loops (PLLs) to generate the desired frequency set. As a rule, indirect synthesizers are not as quick to switch frequencies, but are easier to implement.

Figure 14 shows a block diagram of an indirect frequency synthesizer. Those familiar with PLLs will immediately recognize the structure. The reference frequency, f_I, is related to the output frequency, f_j, by

$$f_j = n_j \times f_I$$

since the VCO and the feedback loop forces f_j/n_j to equal f_I. As can be seen in **table 4**, indirect frequency synthesizers, manufactured by several IC companies, can produce output frequencies above 100 MHz. To achieve greater frequencies, multiplier stages must be added.

The duration of a single frequency hop (t_h) may be longer or shorter than the duration of a message bit. If the message bit duration is longer than t_h, then the system is called a *fast hop FH* system since the hop rate is greater than the message bit rate. Otherwise, the system is termed *slow hop FH*. The advantage of a fast hop system is that if there is interference on one of the hop frequencies, the garbled message bit may still be present at the next hop frequency. For slow hop systems, error correction coding is needed since QRM on a given hop frequency could obliterate several message bits.

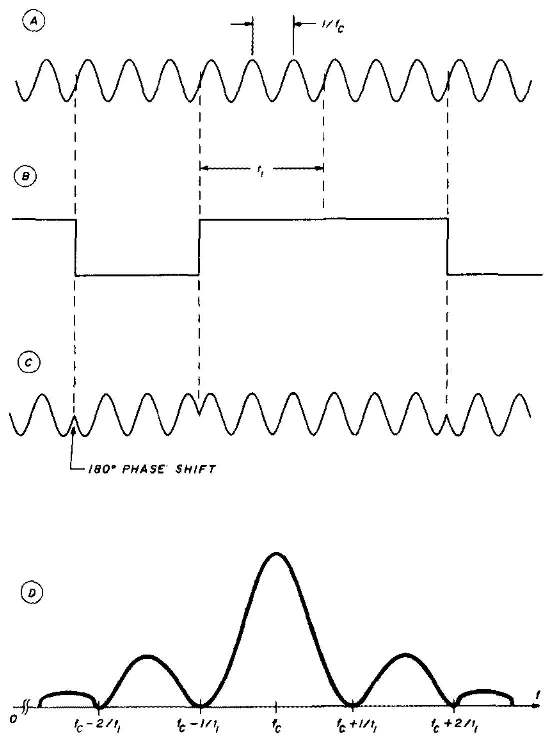


fig. 12. Double-balanced mixer signals: (*A*) RF carrier input, (*B*) digital input, (*C*) output signal, and (*D*) output signal-frequency domain.

Reception of an FH-SS signal is achieved by synchronizing the receiver frequency with that of the transmitter's. Once synchronization occurs, the receiver output is identical to FSK, with the mark and space separated by f_1 Hz. The demodulator can consist of band pass filters which are compared to determine the value of the message bit. The actual message

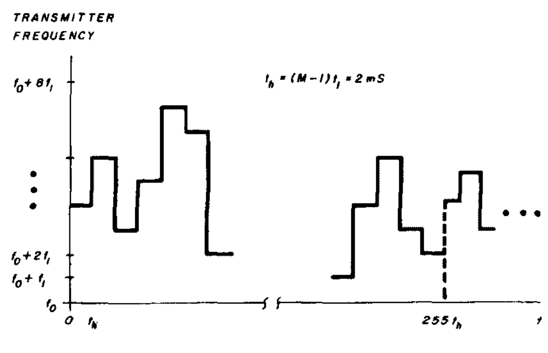


fig. 13. Frequency hopped signal for M = 3, m = 9, t_1 = 1 ms.

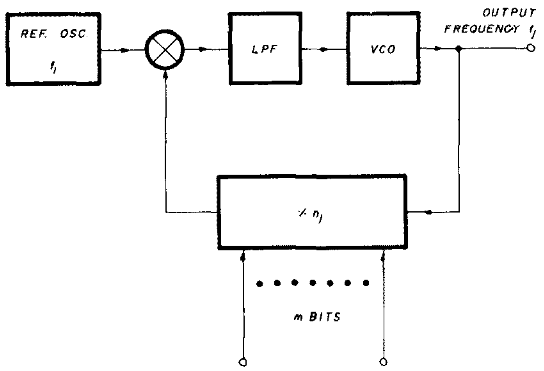


fig. 14. Indirect frequency synthesizer.

is then reconstructed by using a D/A and a low-pass filter.

Synchronization must be acquired in roughly the same manner as in the DS case. A computer shifts the M-1 bit PN code until the receiver tracks the transmitter through each hop. As cited in the DS case, a shorter PN code period ensures quicker acquisition of the transmitted signal.

With indirect synthesizer chips and MSRG chips readily available, it appears that FH systems could be constructed with slight modification to existing VHF/UHF gear. Certain standards such as hop duration, frequency band allocation, and the number of hop frequencies need to be developed, however.

future of spread spectrum

In late May, 1985, the FCC approved docket 81414, which allows the use of spread spectrum on all ham frequencies above 420 MHz as of May, 1986. Temporary authorizations are now being given to those Amateurs interested in experimentation.

In less than six months from now, we'll have a new mode of communication unlike any other we've ever tried. With this new mode, our hobby may take a big step toward reaching state-of-the-art digital communication techniques. There's a lot of work to be done, though; defining protocols for Amateur SS will not be simple.

conclusion

The world of digital communications is a new, exciting technological field, and the future is being shaped daily by advances in this area. As Amateur operators, part of our charter is *to increase the reservoir of electronics experts*. While we don't have to be experts on digital communications, it's probably good for us to know the how's and why's of what is going on around us. Perhaps this article has shed some light on a subject that, as timely as it is, has not been widely discussed in the Amateur Radio literature.

For those interested in learning more about spread spectrum systems, the definitive reference is *Spread Spectrum Systems*, by Robert C. Dixon.[17] Dixon's book was the first on the topic, and has recently been revised to include discussions about practical hardware considerations. Also good is a book just released: *Modern Communications and Spread Spectrum* by C.D. McGillem and G.R. Cooper.[13] This book treats practically every type of modulation method and highlights some of the more important concepts of SS communication.

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ham radio

a packet radio primer

General information to get you started

Packet is the most exciting thing to hit ham radio since voice communication. It far overshadows SSB in importance, and it will, I believe, have even more impact on Amateur Radio than the proliferation of repeaters in the early 1970s.

At its simplest, packet radio resembles radio teletype. But there are differences between the two. First, a packet message is not transmitted as it is being typed. Instead, the characters are stored in a buffer and then sent in a block at the transmission speed of the link — up to 5,600 characters per second (cps) at 220 MHz and higher frequencies, 1960 cps for 6 and 2 meters, 120 cps on 10 meters, and 30 cps for the low bands. Thus, even in a hot-and-heavy QSO, transmit duty cycle and channel utilization are low.

Second, each packet station "knows" its own call and recognizes the messages addressed to it. A number of QSOs can occur on one channel simultaneously, yet each station in the connected mode* will have a screen clear of all messages other than its own contact.

Third, because of computerized error checking, you'll see only perfect, noise-and-garble-free transmissions (note that I didn't say "error-free"; I have a problem with my typing) unless you disable the error-checking function. And that's just for starters!

digital repeaters

Packets can be repeatered. If you don't value the friendship of your fellow repeater users, packets can be put through your local 34-94 machine, but the raucous buzz will drive the control operator, and anyone else monitoring, mad. The better method is to use a digital repeater or "digipeater." This is a very simple device — just a regular packet station. The scheme is quite different from a voice repeater. The digipeater receives the packet signal, stores it in digital form and checks it for errors. If there are no errors, the digipeater retransmits it.

All operation is on one frequency — no duplexers are required, receiver desensitization by the accompanying transmitter cannot occur, and each packet is checked for errors at each digipeater. Unlike voice transmissions via repeaters, packet messages can be sent long distances on the VHF/UHF bands by naming a number of sequential digipeaters to form a path to the destination.

Although, at this moment, most packet activity takes place on 2 meters, there are a number of stations on both coasts operating "gateways," low band packet stations designed for long haul message passing with collection and distribution at either end via VHF or UHF.

packet applications

The four main uses for packet radio at the moment appear to be the following:

- **normal, rag-chew type QSOs.** This of course, includes the exchange of any type of traffic between two stations;
- **direct message** passing (if the destination station is running, I can leave a message on it, without the help or intervention of the operator);
- **Packet Bulletin Board Systems (PBBSs)** similar to the telephone-accessed bulletin board/program exchanges used by computer hobbyists, and **packet mail boxes**, which are usually operated in conjunction with a PBBS. In these, a message can be left for a specific ham by call, for a group such as GLBers, or for everybody (in this case, the call entered as "ALL"). Stations operating mail boxes usually transmit the calls for which they hold traffic. Thus it is not necessary to "check into" the PBBS to know if you have mail.

computer bulletin boards

It's not necessary for your QSO to be with a real, live human. Back in the middle 1970s, several computer operators in the Chicago area set up Remote CP/M (RCP/M) computers with personal message services, bulletin boards, and facilities to exchange computer programs, accessible to anyone with a telephone modem and a teletype or other computer terminal.

These "tele-computing" facilities have proliferated, and it's a poor town, indeed, that doesn't have at least two or three telephone-accessed, computerized bulletin board program exchanges devoted to some com-

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*"Packetese" for being in QSO —Ed.

puter-related or other special interest. (The problem with these marvelous facilities is that if you really get into them, your telephone bill approaches infinity asymptotically.)

This type of activity is a natural for packet radio, and packet bulletin board systems (PBBSs), usually running CP/M, are now springing up nearly everywhere. Many of these stations, based on surplus Xerox 820 computer boards, use software donated to the public domain by Hank Oredson, WØRLI, and distributed by ARRL and through the Newington, Connecticut, FIDO-Net bulletin board (203 665-1114). These bulletin boards offer the advantages of their telephone counterparts without running up your phone bill, and offer an additional convenience: most have a "beacon," a short automated transmission announcing their presence at regular intervals.

These PBBS beacons often include all ham calls for whom the board presently holds messages. Thus, unlike the telephone-accessed boards, you need not "log on" to know if you have mail; just monitor the channel and watch the beacon. In this application, the old ham expression of "reading the mail" takes on a new significance.

for the future

There are two precursors of things to come. The first is the concept of a "local user." This means that a ham tells area PBBS operators which PBBS he considers "home." Messages left for him on other boards are then forwarded to his "home" board. This is now handled manually, but automatic forwarding is only a computer program away.

Second, it's now necessary for a packeteer to determine the digipeater string and enter the calls manually. To assist with this, many PBBSs carry area system maps showing digipeater calls and station locations (an abbreviated map is shown in **fig. 1**). However, as I write, a number of hams are working on computer programs that monitor digipeater traffic, picking up routes and maintaining a dynamic area map in real time.

It doesn't take a great leap of imagination to see where all this is leading: automatic path selection and dynamic call forwarding, both "transparent to the user"; all you do is type the call and the computer and packet board does the rest (with a name/call file for your friends, all you'd need to type would be the name). We're rapidly approaching the point reached years ago by television's Napoleon Solo, who, when stranded on a remote Pacific isle, simply whispered "Open channel D," into his fountain pen. Our main unresolved technical challenge will be to place a workable (and comfortable) keyboard on the side of the pen!

There is one thorn in this rosy future: channel space. I've said that packet transmission is error-free, but I carefully avoided any reference to transmission times. At the moment, channels are relatively quiet, and over short paths things can happen quickly. For example, working through two 2-meter digipeaters handling little or no other traffic, packet delivery times run on the order of 4 to 5 seconds each. This is quite reasonable. But, add a few more QSOs, a longer path (several digipeaters), or someone using a PBBS (which shovels out long program or message packets as fast as it can), and things bog down quickly.

While we're on the subject of transmission speeds, two additional notes are necessary. First, the speeds given at the beginning of the article are FCC-permitted *maxima*. While 300 Baud (30 cps), the legal maximum, is used on the low bands, on 2 meters the universally used speed is presently 1200 Baud (120 cps) rather than the allowed 19.6 kilo-Baud, purely because the 1200 will go through a normal voice channel while the higher data rates (which require greater bandwidths) will quickly run aground in the intermediate frequency amplifier and audio circuitry of a normal voice rig.

Second, 1200 Baud sounds like a nice, snappy exchange rate, particularly if you're accustomed to a 300 Baud terminal. Sorry about that; the 1200 Baud is the character transmission rate. When it comes to actual message or data exchanges, the through-put (useful traffic passed) will be at a much lower effective speed, especially if noise or a busy channel forces repeats of transmissions. This is due to overhead within the packet (message header and such) and the necessity of getting acknowledgement for each packet.

Of course, packet activity has no way to go but up as more people get involved. Again, we have technical solutions that will (we hope) be here before the problems are — inexpensive "ham type" digipeaters up at UHF and microwave where transmission speeds can be increased dramatically. This doesn't mean you'll have to put a 1296 MHz rig in your car; local traffic can still be handled on 2 meters, but the heavy and long-distance communications can go on the microwave links.

equipment

You need only three pieces of equipment to operate on packet radio: a 2-meter transceiver (nothing special, that old rock-bound clunker in the basement will probably do, with some tuning); a Terminal Node Controller (TNC — the actual "packet board"); and a computer or computer terminal with which to communicate with the TNC.

Of course, there are always other ways. If you want to save time and reduce aggravation, there's the Pack-

fig. 1. At right is a graphic representation of Packet Radio links believed to exist on 145.010 MHz on the East coast of North America. The area covers part of Canada and extends to North Carolina. Another file, called SOUTMAP.NNN, covers the area from Virginia south to Florida.

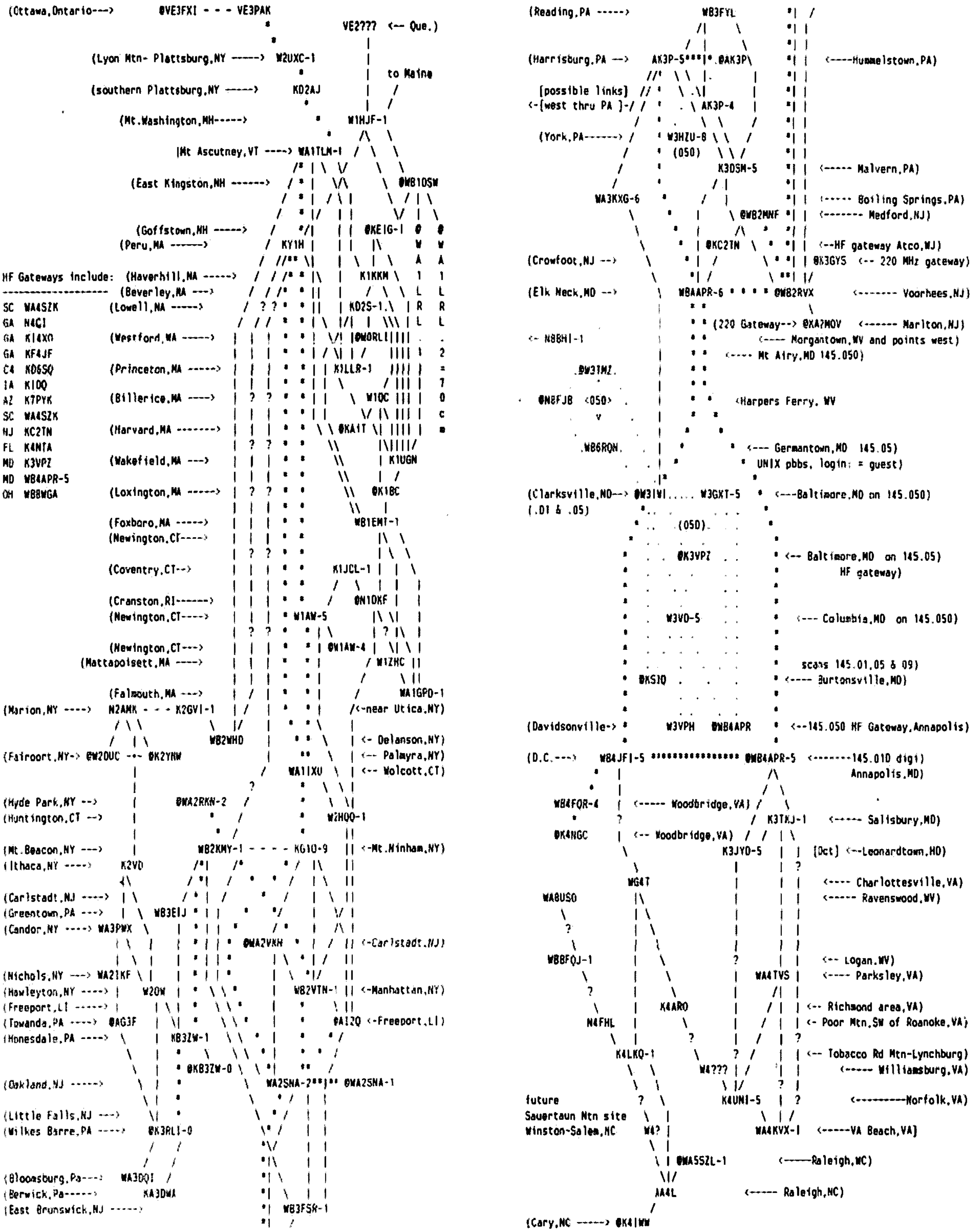
Included in these maps are digipeaters, Packet Bulletin Board Stations (PBBS), and home stations usually left on for digipeating. Updated maps covering this and other areas are available from stations running WØRLI-compatible software for PBBSs, mailboxes and/or gateways; these stations are indicated by the use of the "@" character preceding the call sign.

The primary routes used for mail forwarding in EASTNET are marked with asterisks (****), while other links are shown by either | \ or / as connecting characters. All "****" routes are presently on 145.010. In the Baltimore/Washington area .050 is heavily used for local traffic. Links on 145.050 MHz in Maryland, West Virginia, and Pennsylvania are connected on this map by dots. 220 MHz trunk links will parallel 145.010 in the future.

To improve readability of this linking you might want to use felt tipped pens to outline the various linking paths. Links marked with a question mark indicate a link of unknown reliability. If any links shown on this map prove to be unreliable or nonexistent, please drop me a note. Send the info to me, K1HTV @ W3IWI, via one of the many auto-forwarding PBBBs or via the U.S. mail (see Callbook for address).

Thanks to Rick Zwerko, K1HTV, one of 10 acting directors of the Mid-Atlantic Packet Radio Council (MAPRC). Formerly Vice-President for Operations, AMSAT, Rick also served as a member of the AMSAT Board of Directors and was past president of the Northeast VHF Association.

— WA1FHB



(Ottawa,Ontario---> @VE3FXI - - - VE3PAK
VE2??? <-- Que.)
(Lyon Mtn- Plattsburg,NY -----> W2UXC-1
to Maine
(southern Plattsburg,NY -----> KD2AJ
(Mt.Washington,NH-----> W1HJF-1
(Mt Ascutney,VT ----> WA1TLN-1
(East Kingston,NH -------> @WB1DSW
(Goffstown,NH -----> @KE1G-1
(Peru,MA -------> KY1H
HF Gateways include: (Haverhill,MA -----> K1KKM
(Beverley,MA --->
SC WA4SZK (Lowell,MA -----> KD2S-1
GA N4CJ
GA KI4XO (Westford,MA -----> @W0RLI
GA KF4JF
C4 KD6SQ (Princeton,MA -----> K1LLR-1
IA K1DQ
AZ K7PYK (Billerica,MA ----> W1OC
SC WA4SZK
NJ KC2TN (Harvard,MA --------> @KA1T
FL K4NTA
MD K3VPZ (Wakefield,MA ---> K1UGN
MD WB4APR-5
OH WB8WGA (Loxington,MA -----> @K1BC
(Foxboro,MA -----> WB1EMT-1
(Newington,CT---->
(Coventry,CT--> K1JCL-1
(Cranston,RI------> @N1DKF
(Newington,CT----> W1AW-5
(Newington,CT---> @W1AW-4
(Mattapoisett,MA ----> W1ZHC
(Falmouth,MA ---> WA1GPD-1
(Marion,NY ----> N2AMK - - - K2GVI-1 (<-near Utica,NY)
WB2WHD (<- Delanson,NY)
(Fairport,NY-> @W2DUC --- @K2YNW (<-- Palmyra,NY)
WA1IXU (<-- Wolcott,CT)
(Hyde Park,NY --> @WA2RKN-2
(Huntington,CT --> W2HQO-1
(Mt.Beacon,NY ---> WB2KMY-1 - - - - KG1O-9 (<-Mt.Ninham,NY)
(Ithaca,NY ----> K2VD
(Carlstadt,NJ ---->
(Greentown,PA ---> WB3EIJ
(Candor,NY ----> WA3PWX
@WA2VXH (<-Carlstadt,NJ)
(Nichols,NY ---> WA2IKF
(Hawleyton,NY ----> W2OW WB2VTN-1 (<-Manhattan,NY)
(Freeport,LI ----->
(Towanda,PA ----> @AG3F @AI2Q <-Freeport,LI)
(Honesdale,PA ----> KB3ZW-1
@KB3ZW-0
(Oakland,NJ -----> WA2SNA-2 @WA2SNA-1
(Little Falls,NJ --->
(Wilkes Barre,PA ----> @K3RLI-0
(Bloomsburg,Pa---> WA3DQI
(Berwick,Pa-----> KA3DWA
(East Brunswick,NJ -----> WB3FSR-1
(Reading,PA -----> WB3FYL
(Harrisburg,PA ---> AK3P-5 @AK3P (<----Hummelstown,PA)
[possible links]
<-[west thru PA]- AK3P-4
(York,PA------> W3HZU-8
(050)
K3DSM-5 (<----- Malvern,PA)
WA3KXG-6 (<----- Boiling Springs,PA)
@WB2MNF (<------- Medford,NJ)
@KC2TN (<--HF gateway Atco,NJ)
(Crowfoot,NJ --> @K3GYS <-- 220 MHz gateway)
(Elk Neck,MD --> WB4APR-6 @WB2RVX <------- Voorhees,NJ)
(220 Gateway--> @KA7MOV <------ Marlton,NJ)
<- N8BHI-1 <---- Morgantown,WV and points west)
<---- Mt Airy,MD 145.050)
@W3IMZ
@N8FJB <050> <Harpers Ferry, WV
WB6RQN <--- Germantown,MD 145.05)
UNIX pbbs, login: = guest)
(Clarksville,MD--> @W3IWI W3GXT-5 <---Baltimore,MD on 145.050)
(.01 & .05)
(050)
@K3VPZ <-- Baltimore,MD on 145.05)
HF gateway)
W3VD-5 <--- Columbia,MD on 145.050)
scans 145.01,05 & 09)
@KS3Q <---- Burtonsville,MD)
(Davidsonville-> W3VPH @WB4APR <--145.050 HF Gateway,Annapolis)
(D.C.---> WB4JFI-5 @WB4APR-5 <-------145.010 digi)
Annapolis,MD)
WB4FQR-4 (<----- Woodbridge,VA)
K3TKJ-1 <----- Salisbury,MD)
@K4NGC (<-- Woodbridge,VA)
K3JYD-5 [Oct] <--Leonardtown,MD)
WG4T <---- Charlottesville,VA)
WA8USO <----- Ravenswood,WV)
WB8FQJ-1 <-- Logan,WV)
WA4TVS <---- Parksley,VA)
K4ARO <-- Richmond area,VA)
N4FHL <-- Poor Mtn,SW of Roanoke,VA)
K4LKQ-1 <-- Tobacco Rd Mtn-Lynchburg)
W4??? <----- Williamsburg,VA)
future K4UNI-5 <---------Norfolk,VA)
Sauertaun Mtn site
Winston-Salem,NC W4? WA4KVX-1 <-----VA Beach,VA]
@WA5SZL-1 <-----Raleigh,NC)
AA4L <----- Raleigh,NC)
(Cary,NC -----> @K4IWW

eterm, a portable unit ($995) that combines a TNC and a computer terminal in one sleek-looking designer case (see photo). If you go that route, you can disregard the rest of this article and be on the air half an hour after unpacking the box . . .

I claim no special expertise in the area of 2-meter transceivers. I'm using an Azden PCS4000 because I happen to have it, and it's working fine. The only caveat is that most 2-meter packet activity is below 146 MHz and some of the older narrow-band equipment needs a tweak to get down there.

I'm using a GLB PK-1 Terminal Node Controller. I chose it because it appears to be the least expensive one available ($165). At 4-1/2 x 9-1/2 inches (11.4 x 24 cm) it's also the smallest one I've seen. It also requires only a single-voltage power supply (+ 12 VDC at 170 mA) and it doesn't mind "mobile-type" voltage excursions; I've used it from 14-1/2 to 9 volts without a hiccup. And, like other TNCs on the market, the Z-80-based GLB features "dynamic programming"; the manufacturer frequently releases a new ROM offering enhanced and improved features.

Other TNCs are available from Vancouver Amateur Digital Communications Group, Tucson Amateur Packet Radio, Heath, AEA, Kantronics, Ashby, and Packeterm, to name a few (**fig. 2**). Richardson software converts a TRS-80 Model 1, 3, 4, or 4P into a computer/TNC*.

The last necessity, the terminal, offers the most opportunity for self-expression. It can be anything from a Model 15 Teletype to a microcomputer such as a Sinclair ZX-81 or a Commodore 64. The microcomputer route is the most popular. Of course the micro must have a serial port (or an adapter to provide one) as well as a modem program. (The Model 15 Teletype was preprogrammed at the factory.)

getting started

As with everything else, the most difficult part is getting started, especially if you don't have a packet Elmer around. You'll need either a reasonably local packet station that will give you a strong signal or two set-ups that you can work back-to-back. An independent monitor receiver is a big help.

There's nothing special about the monitor receiver. It can be any kind of tunable or fixed-frequency rig capable of receiving your area's packet channel(s). Be aware that some of the programmable (no-crystal) scanners won't tune lower than 146 MHz without special measures. For example, on my 16-channel Regency "Touch," Model ACT-T-16K, I must press, in order, MA, 9, CL, PR, then key in the frequency I want, and hit PR again.

Synchronous Packet Radio Using the Software Approach – AX.25, by R.M. Richardson, W4UCH, available from Ham Radio's Bookstore, Greenville, NH 03048 ($21.95 plus $3.50 postage & handling)

fig. 2A. Cynthia "Sam" Hensley, KY1D, operates Packeterm™ IPT, a portable terminal and TNC combined in one unit, from summit of Pitcher Mountain in Stoddard, New Hampshire.

This receiver is then set up on its own 1/4-wave whip near your packet station. Because a TNC requires a better signal to noise ratio than voice, if another station sounds reasonable on the quarter-wave whip, it will probably be fine for packet, assuming that you are using an outdoor gain antenna for the packet transceiver.

The monitor is used, first, to compare your station's deviation with others in the neighborhood, and, second, to keep track of when (and how) your station is transmitting. After your packet set-up is thoroughly established and proven, you may wish to return this radio to its prior service monitoring the local police, but till then it'll be invaluable in getting you started.

To begin with, you have to get your computer or terminal working with your TNC. This requires, first, an RS-232 interface from the computer or terminal to be connected with the RS-232 interface on the TNC, and, second, if you're using a computer, driver software (usually a "Modem" or "Terminal Emulator" program) to access the RS-232 port. Because the programming depends on the brand and type of computer used, I'll leave that part to you and your software dealer or local computer guru.

I can, however, provide some advice on using the RS-232.

the RS-232

First, the actual RS-232 specification doesn't define a physical connector, although a DB-25 is usually used. Second, RS-232 defines interfaces for modems and for terminals, but not for computers. Third, two

identical RS-232 interfaces (two modem interfaced or two terminal interfaces) won't work together because both will be transmitting data on one line and both will be "listening" on another line. Therefore, before you actually hook things together, compare the instructions for your computer and your TNC.

Normally, the computer will talk (transmit data, data out, or TxD) on pin 2 of the DB-25 and the TNC will listen (receive data, data in, or RxD) on the same pin number. If your literature shows that your two devices work this way, use a pin-for-pin cable. If, on the other hand, both units are talking on the same line, you'll have to swap pins 2 and 3 at just one end of the cable. (If you make the swap at both ends, you're back where you started!) Along with that swap, go swaps on several of the control lines, although ground *always* remains on pins 1 and/or 7. Check your literature.

If you find that you do require some line swaps, but

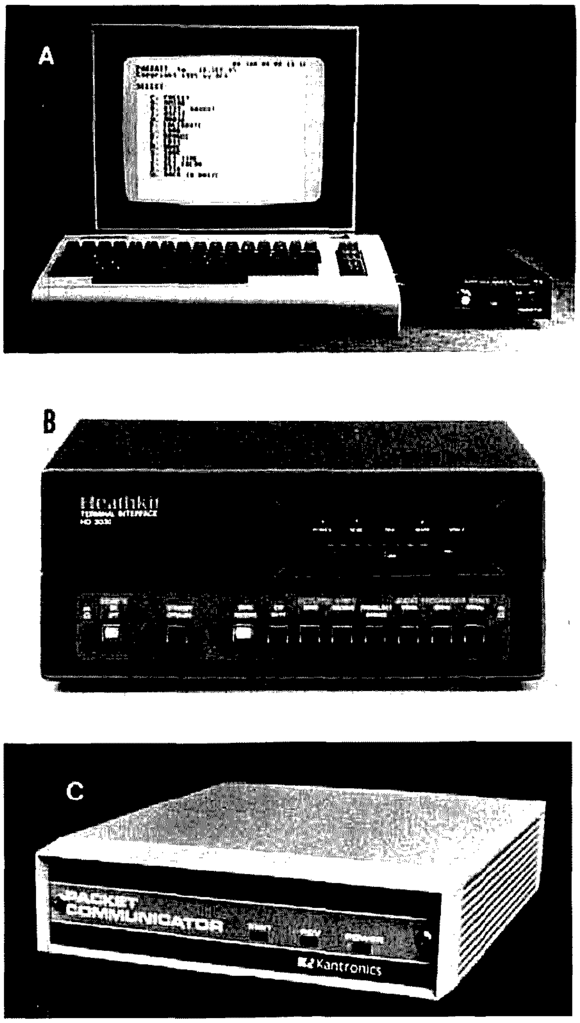


fig. 2B. Packet terminal units from (A) AEA, (B) Heathkit, and (C) Kantronics.

all you have is a pin-for-pin cable, check its connector type. There are three types of DB-25 connectors in common use commercially: insulation displacement connector (IDC), crimped onto ribbon cable; solder cup connector with wire leads soldered in as required; and crimp-type, where the pins are crimped onto the wires and then popped into the connector shell.

If you have only a ribbon cable, you're stuck. With the solder cup connector, changing conductors around is relatively simple, assuming that you have both patience and a small soldering pencil. Changing the pins around on the crimp-type connector is even easier than resoldering if you can get the little plastic insertion and extraction tools. Both tools (which slip together for storage) are slotted lengthwise for slipping over the wire. The tool is then slid down the wire so that its point enters the back of the connector block around the connector pin. Gently pullng the wire slides the pin out of the block. Reinsertion is just the reverse. Place the tool on the wire up against the pin and use the tool to seat the pin in the block.

Although these tools are inexpensive, they may be difficult to find. Made of plastic, they will break in time. You might try purchasing several from an industrial distributor who stocks the connectors.

listening to the data

If you're sure the wiring is under control but the interface still doesn't work, check to make certain that both units are really transmitting data when they should. Look at the TxD line from the computer. (An oscilloscope is ideal for this but you can use the audio channel of your video monitor, if it has one, or another audio amplifier and speaker.

Remove the plug cover from the cable connector at either end. Make sure that the audio channel ground is connected to the computer system ground. With the volume of the audio channel set low (to avoid ear-shattering surprises) check the audio by touching the exposed audio connection with your finger while the rest of your body is ungrounded. You should get a loud AC hum. Then, using a small-gauge conductor wedged into the back of the DB-25, make contact between either pin 2 or pin 3 and the audio input.

Now try a command to the TNC. With the GLB PK-1, the first thing it wants to see is a carriage return <cr> to establish the baud rate, which should be 9600 Baud or slower. (The baud rate between TNC and "terminal" has nothing to do with the "on the air" data rate of 1200 Baud.) When the GLB receives its carriage return, it responds with a dozen-character sign-on. If serial data is present on the line you've chosen, you'll get a raucous buzz (or a short burp) for a single character like the <cr>. Try this a couple of times. If the power doesn't come on cleanly (i.e., bounce in the power switch) it can discombobulate

the GLB's initialization routine. Hit GLB reset and try <cr> again.

If you still don't get the buzz, the most likely cause is that a required RS-232 control signal is not being handled properly. Of course, this requires a check of the RS-232 specifications as interpreted by the two particular devices that are giving you grief. For example, with the GLB PK-1, check to see that the computer or terminal is putting a high (>3VDC) on RTS (Ready To Send) (DB-25 pin 4 on the GLB), and that the computer doesn't need more control signals than the high that the GLB puts on its CTS (Clear To Send) (DB-25 pin 5) line.

If either the computer or the TNC is not getting the control signal(s) it requires, you won't hear data on either line from the end(s) with the problem. Check your control lines with a VOM.

Once the terminal-TNC interface is working, you are ready to hook the TNC to your transceiver, following the TNC manufacturer's guidelines.

Now see if you can receive. Connect all the equipment up and turn it on, instructing your packet board to display everything without checking the packets for transmission errors ("Garbage mode" on the GLB, **SG-E**). Wait for activity on the channel (as indicated by your transceiver's S-meter or the monitor radio) and see if it prints. (Very short transmissions may be connect requests or acknowledgements that will not yield a printable message.) If that works, try transmitting. Send anything and compare the sound of your transmitter through the monitor radio with the sound of another packet station. *If your signal doesn't sound raspy and disagreeable, your audio is set too low.*

talking to yourself — by radio

If the audio sounds okay, try talking to yourself through the other station. Program your call as the destination as well as the originator, making sure that the SSID numbers (usually zero) are the same. Then type in the call of the other packet station or digipeater, again watching the SSID. Many digipeaters use an SSID of one, with zero used by the trustee's home station. W1AW has several packet stations with SSIDs running up to 5! With the GLB your destination is set with **SD** and your digipeater(s) with **SV** (send via).

Now issue the command to connect (**AC** on the GLB). Your transmitter should come on for a short period (less than a second) followed almost instantly by a similar-length transmission by the other station. Your terminal should then "ring its bell" and display **-Connected to <your call>**. Now type a short message and hit your "dispatch" character (on the GLB this defaults to a line feed). The message should be duplicated almost immediately on your screen, then be followed in a second or two by another bell and an **-Ack**.

The **-Ack** means that the "receiving-you" has acknowledged the message back to the "sending-you." If the **-Ack** is not received by the TNC, the message will be held in the TNC's buffer and the transmission repeated. With the GLB, to disconnect, type **Control-C**. When disconnect is complete, the screen will show **#1**.

If all that worked, you're ready for a real, live QSO or a longer path test. On multi-hop self-connects, you must provide the *entire* round trip in the digipeater string. (Like any other computer, the TNC is wonderously fast but very stupid.) Thus to self-connect through W1AA-1, W2BB-0, and W3CC-1, for example, your string must read W1AA 1, W2BB 0, W3CC 1, W2BB 0, W1AA 1.

If all this works, you're home free. If questions or problems arise, you can connect with another packeteer in your area and ask for help. You can also access your local PBBSs. Start with the one that's on the shortest, quietest, most reliable path until you get the hang of it. (The wee, small hours of the morning are the best time for this experimenting.) All you have to do is connect with the PBBS station, then be patient while it announces itself and gives you its prompt line ending with CP/M's >. It will then be looking for a letter command followed by a carriage return (and on the GLB, a line feed to send the packet). For starters, try an "H" (for Help). It should reward you with a list of its commands and explanations. If you have a printer, turn it on so you'll have a hard copy for future reference.

Of course, there's a great deal more to packet radio than I've mentioned here, and the field is changing rapidly, but, as the Chinese say, "The longest journey begins with but a single step." As with any Amateur Radio activity, the biggest and most important step is just getting on the air so you can contact fellow hams. I trust that you'll enjoy the rest of this unending journey.

ham radio

automatic frequency and deviation tester for packet radio

Measure your packet signals from the comfort of your home

Recent observations of packet radio signals revealed some startling facts. Several stations intending to be on 145.01 MHz were found as much as 3 or 4 kHz off frequency. Deviation levels also varied quite a bit.

Off-frequency operation and overdeviation cause distortion of the audio signal. Too little deviation results in a poor signal to noise ratio. In all these cases, the result is the same: a modem is less likely to demodulate the signal properly. Packets are retransmitted an excessive number of times and the channel gets clogged up.

The New England Packet Radio Association tried to improve the situation by having a calibration session at one of the regular meetings. A frequency counter, deviation meter, and qualified engineers were present to make sure everybody's equipment was properly adjusted. There was plenty of advance publicity, but no more than a few people bothered to bring their equipment.

Several months later a new beacon appeared on the air, transmitting *"WB2OSZ>BEACON: frequency/deviation tester available."* During the first few days of operation, many stations connected and received the following message:

"Welcome to the WB2OSZ automatic frequency/deviation tester. Instructions:

1. Send several non-blank lines.
2. Wait for reply after each.
3. Ignore any occasional erratic values.
4. Disconnect when done.

Recommended deviation is no more than 3 kHz. Anyone who disagrees with the results is invited to supply a reference more accurate than my 2AT. John"

Each time a packet was sent to the automated station, a reply was sent back in the form, *"Your frequency is about 1.8 kHz too high. Deviation is about 2.6 kHz."*

circuit description

The voltage from the detector of an FM receiver is proportional to the frequency of the incoming radio signal. Extracting the DC component of this will provide a measure of the carrier frequency. The peak amplitude of the AC component is proportional to the deviation. **Figure 1** shows a block diagram of a system designed to extract these parameters.

The first step is to obtain a DC-coupled signal from the demodulator of your FM receiver. The entire signal flow from receiver to computer is shown in **fig. 2A** and the individual circuits detailed in the **figs. 2B** through **2E**. **Figure 2B** contains a circuit used with an old VHF Engineering transceiver; it should work with anything else using an LM 3065, an MC 1358, or a CA 3065 quadrature detector. (This part will have to be customized for your particular rig.) It produces a zero volt output for a carrier on the desired frequency and changes by a half volt for each kHz change.

The next step is to extract DC and AC components of the signal with low- and high-pass filters (**fig. 2C**). U2 and associated components form a low-pass filter. Their output is proportional to the amount the incoming signal is off frequency. For example, −1.5 volts means the signal is 3 kHz too low.

The AC component, of course, is the audio. It is extracted by a high-pass filter composed of C3 and R12. This is fed through a full wave rectifier, peak detector, and another low-pass filter. R14 limits the peak current out of U4. (Without it, nasty spikes appeared on the power supply output). R18 boosts the gain slightly above unity to compensate for a small loss in the high-pass filter and peak detector. The result

By John W. Langner, WB2OSZ, 115 Stedman Street, Chelmsford, Massachusetts 01824

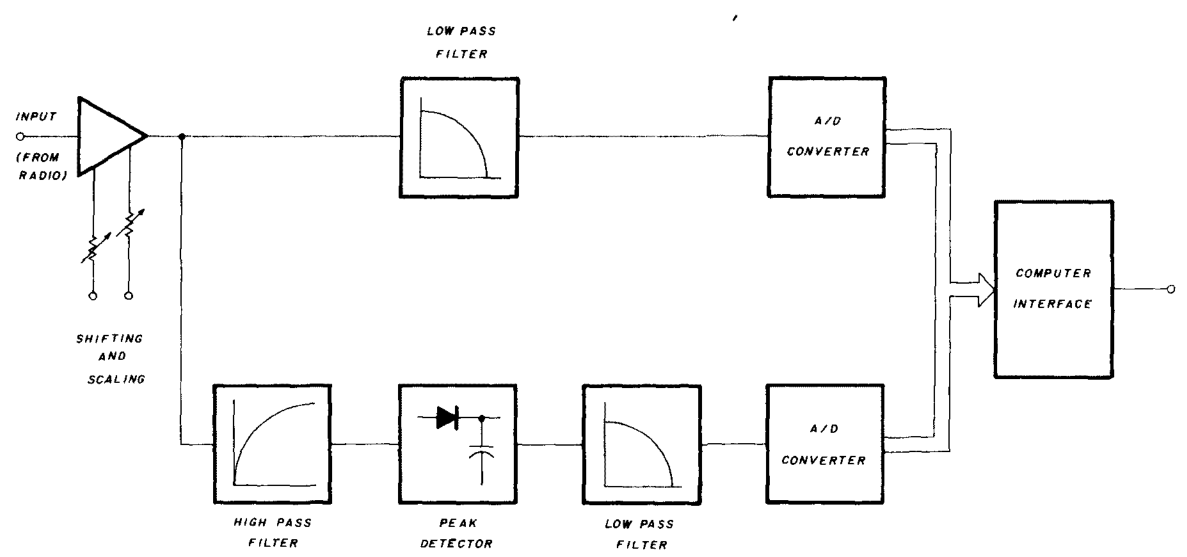


fig. 1. Block diagram of system to determine frequency and deviation of received signal.

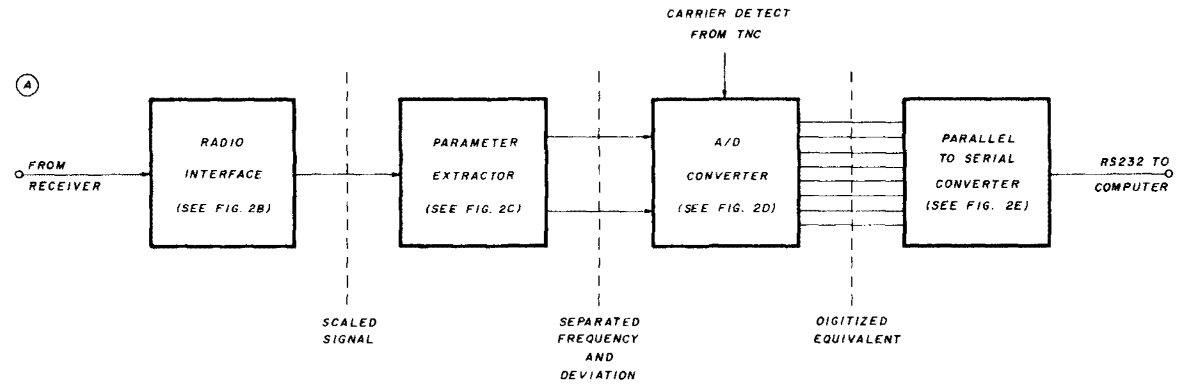


fig. 2A. Radio interface.

table 1. Integrated circuit pin designations for the supply lines.

	ground	+5 volts	−12 volts	+12 volts
555	1	8		
741			4	7
ADC 0804	10	20		
1458			4	8
74121	7	14		
AY-5-1013	3	1	2	

is a voltage proportional to the peak deviation, again 0.5 volt/kHz.

To build a useful piece of test equipment, connect these voltages to a pair of meters. You can also integrate your system with a computer to open up many possibilities for automated operation. If you have a computer with joy stick input, you already have a pair of analog-to-digital converters and don't have to build the rest of the circuit. For those with computers not intended for playing games, additional external circuitry is required.

The analog voltages must be converted to digital signals the computer can understand. This is accomplished by the ADC0804 A/D converters in **fig. 2D**, which convert a voltage in the range of 0 to 5 volts into a corresponding number in the range of 0 to 255 in less than 20 microseconds. U7 and U8 determine when to start the conversion. CR3 through CR6 protect the inputs of U5 and U6 from voltages that could

damage them. CR3 through CR6 should not conduct during normal operation.

A purist may wish to connect a precision 2.5 volt reference to the V_{REF} pins. In its absence, internal voltage dividers use one half of the 5-volt supply.

When should the A/D converters be commanded to start the conversion? You wouldn't want to do it when an audio carrier is first detected because the low-pass filters wouldn't have had time to settle down. And you can't do it after the computer has received the packet contents from the TNC (terminal node controller) because the audio signal is long gone. I decided

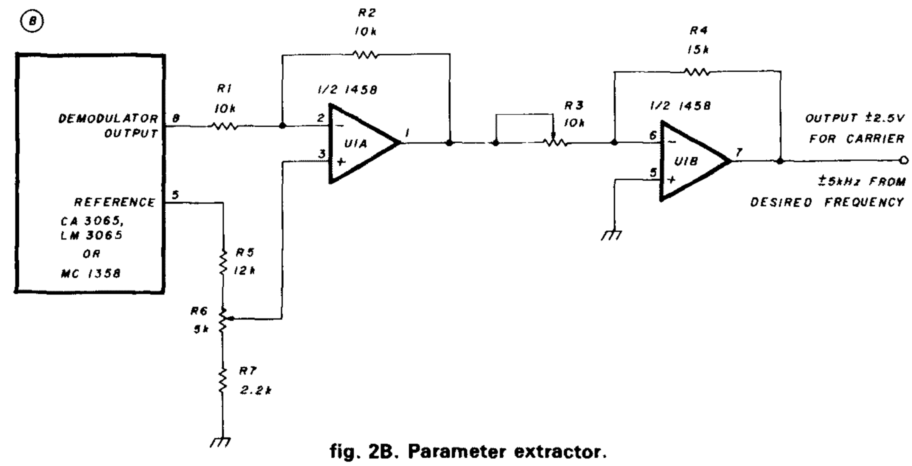


fig. 2B. Parameter extractor.

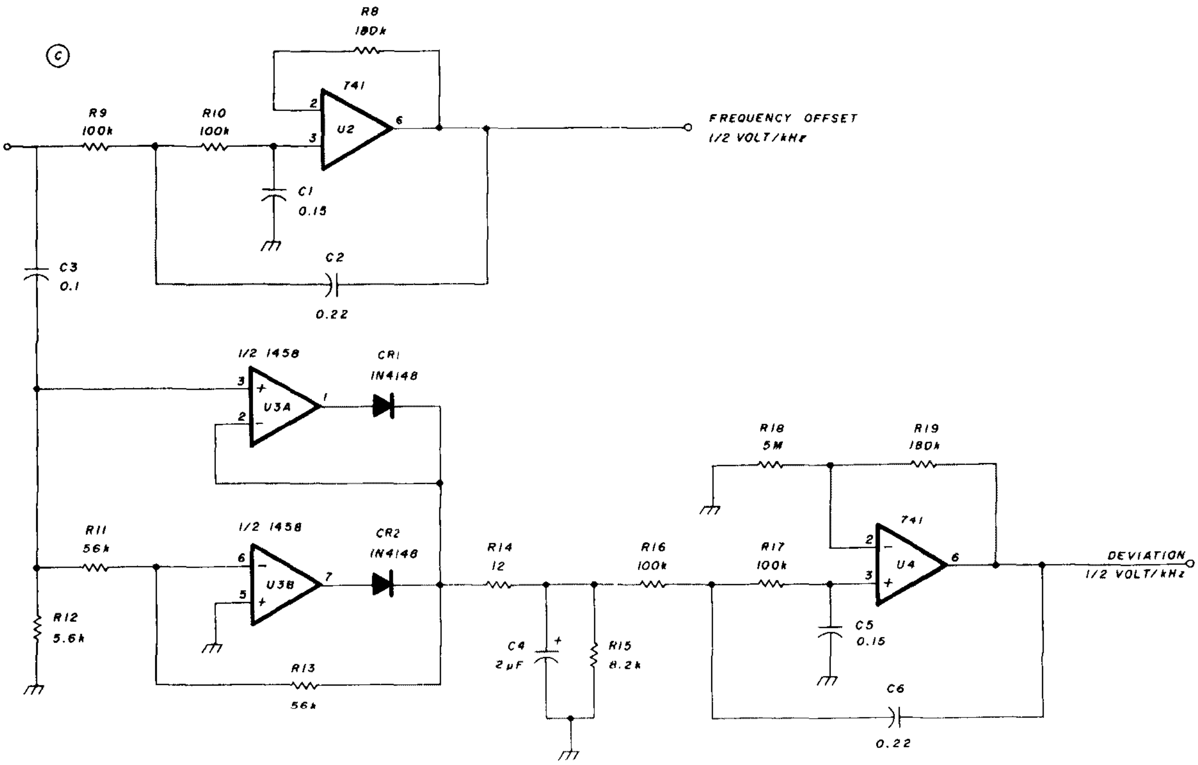


fig. 2C. A/D converter.

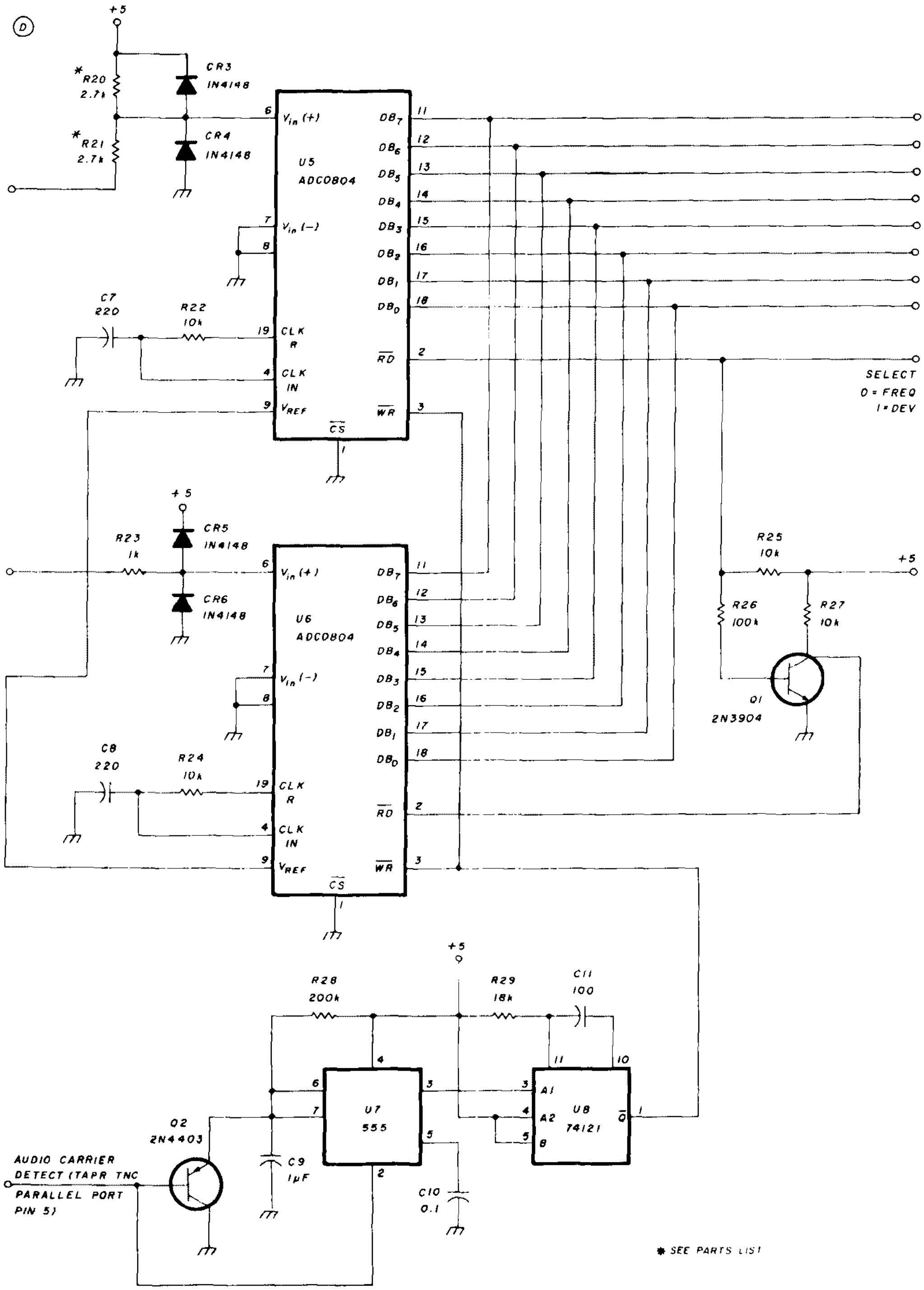


fig. 2D. Parallel to serial converter.

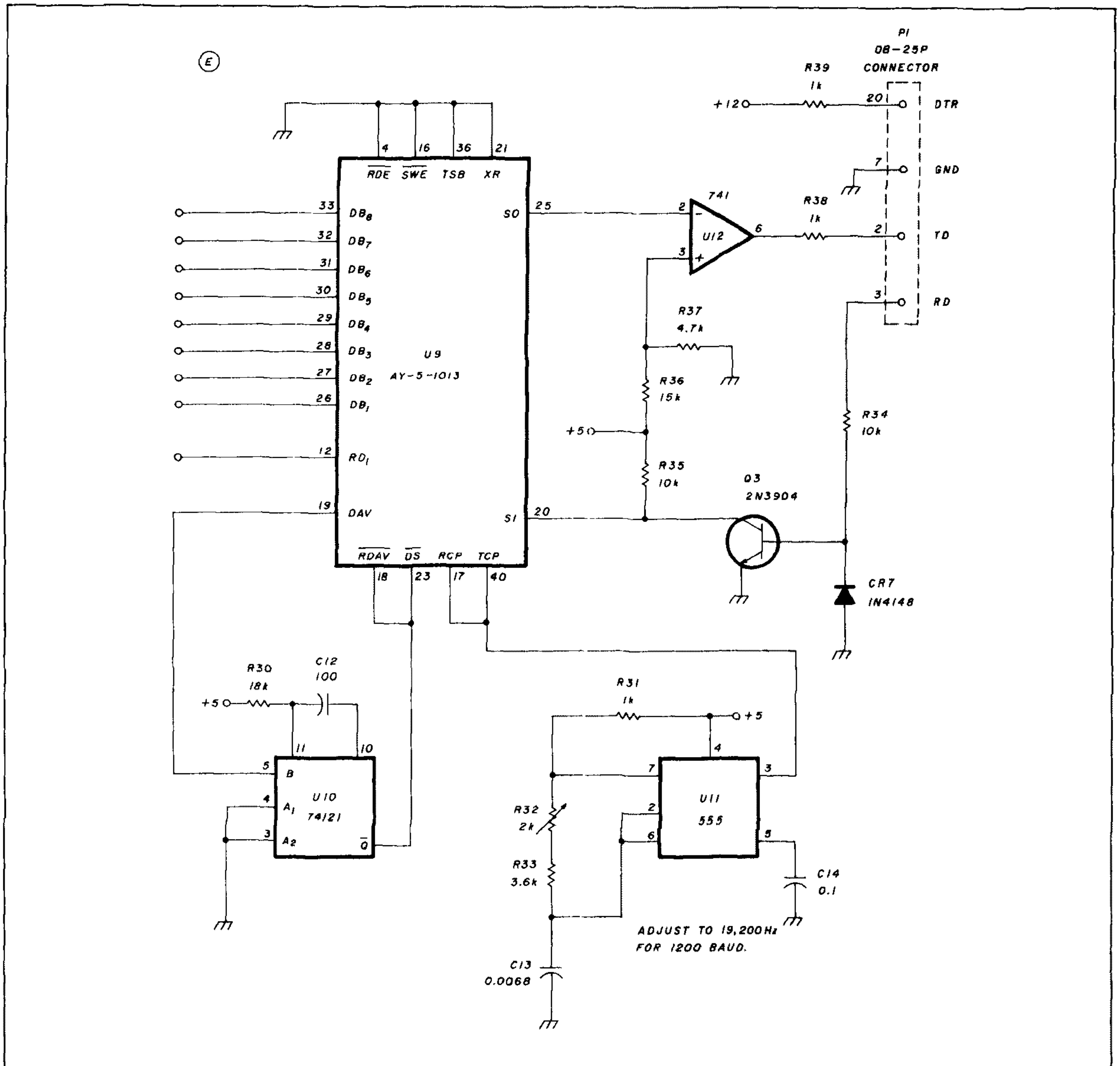


fig. 2E.

to sample the analog signals after the audio carrier had been present for approximately 200 milliseconds. This provides sufficient time for the analog components to settle down but is before the end of the shortest packet. A carrier detect signal is available at pin 5 of the parallel connector on the back of a TAPR TNC.* AEA and Heathkit units, similar to TAPR's, are probably the same.

The computer I'm using doesn't have any parallel input, but does have a spare serial (RS-232) port. The remainder of the circuit converts the parallel data into serial form for communication with the computer as shown in **fig. 2E**.

*Tucson Amateur Packet Radio.

When the computer wants to obtain the most recent measurements, it sends a character to the circuit. The UART (U9) converts the serial character to parallel form and causes U10 to generate a pulse. The least significant bit of the character comes out of pin 12 and selects one of the A/D converters. The pulse from U10 causes the UART to begin conversion of the data from parallel to serial form.

The connections shown for P1 assume a computer port expecting a terminal. If using a modem port, use a DB-25S connector instead and swap the connections to pin 2 and 3.

U11 determines the speed for communications. R32 is adjusted for a frequency 16 times the desired baud rate, for instance 19200 Hz for 1200 baud. The com-

item	description	quantity	approximate cost
C1,C5	0.15 μF mylar (could use 0.1 + 0.047)	2	0.48
C2,C6	0,22 μF mylar	2	0.33
C3,C10,C14	0.1 μF disc ceramic	3	0.15
C4	2 μF electrolytic (could use two 1 μF)		0.34
C7,C8	220 pF mica	2	0.42
C9	1 μF electrolytic		0.17
C11,C12	100 pF mica	2	0.35
C13	0.0068 mylar (could use 0.0047 + 0.0022)		0.24
CR1-CR7	1N4148 or similar	7	0.10
P1	DB-25P connector		2.39
Q1,Q3	2N3904 or similar NPN	2	0.25
Q2	2N4403 or similar PNP		0.25
R1,R2,R22,R24, R25,R27,R34,R35	10 kilohm		
R3	10 kilohm, trim pot		0.29
R4,R36	15 kilohm		
R5	12 kilohm		
R6	5 kilohm, 15 turn trim pot		1.19
R7	2.2 kilohm		
R8,R9	180 kilohm		
R9,R10,R16, R17,R26	100 kilohm		
R11,R13	57 kilohm		
R12	5.6 kilohm		
R14	12 ohms		
R15	8.2 kilhoms		
R18	6 Megohms		
R20,R21	2.7 kilohms. *closely matched		
R23,R31,R38,R39	1 kilohm		
R28	200 kilohms		
R29,R30	18 kilohms		
R32	2 kilohms, 15 turn trim pot		1.19
R33	3.6 kilohms		
R37	4.7 kilohms		
	fixed 1/4 watt resistors	34	0.06
U1,U3	1458, dual op amp	2	0.59
U2,U4,U12	741, op amp	3	0.35
U5,U6	ADC 0804, analog to digital convertor	2	3.49
U7,U11	555, timer	2	0.39
U8,U10	74121, monostable multivibrator	2	0.39
U9	AY-5-1013A, UART		3.95
		Approximate total	$28.00

*not in the catalog; possible substitutions are indicated.

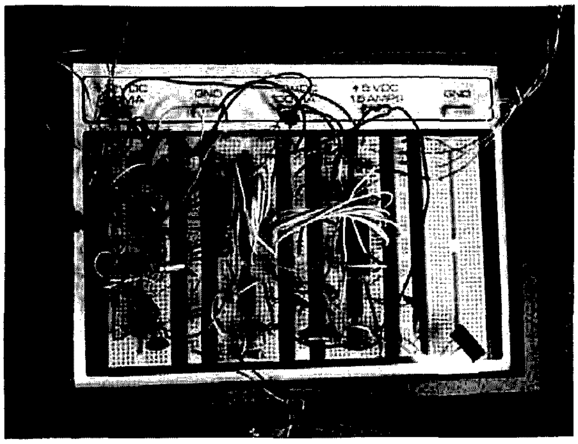


fig. 3. Point-to-point wired circuits.

pleted point-to-point wired circuits are shown in **fig. 3**. Pin designations for the six integrated circuit supply lines are provided in **table 1**.

software description

In the automatic answer mode, the station "advertises" the service available by beaconing and waiting

```
/* Snatch values from A/D */
/* converters before they */
/* have more of a chance to */
/* change. */

n_freq = adc_sample(0);
n_dev = adc_sample(1);

/* Any processing of */
/* received record is now */
/* done. */
/* This could include placing */
/* on screen, printing, */
/* saving in a disk file, */
/* etc. */

        :
        :

/* Process calibration data. */
/* First scale and convert */
/* to decimal digits. */

/* Ignore blank lines. */

if (Rrec_len > 0)
  {
/* Approximate conversions:    */
/* (Extreme values won't get   */
/* thru most receivers.)       */
/*                             */
/* freq    raw   final  text   */
/* offset  data  number result */
/* (kHz)                       */
/* -----   ----  ----   -----  */
/* -10      0    -100  "-10.0" */
/*  -5     64     -50  "-5.0"  */
/*   0    128       0  ".0"    */
/*  +5    192      50  "5.0"   */
/* +10    255     100  "10.0"  */
/*                             */
/* deviation                   */
/* -----                       */
/*   0      0       0  ".0"    */
/*   5    128      50  "5.0"   */
/*  10    255     100  "10.0"  */

   n_freq=((n_freq-128)*100)/128;
   n_dev = (n_dev * 100) / 256;

   for_mat(n_freq, s_freq);
   for_mat(n_dev, s_dev);

/* Package compactly for */
/* screen or log file. */

   if (n_freq > 0)
     strcpy(s_compact, "[ +");
   else
     strcpy(s_compact, "[ ");
```

fig. 4. Segment of a routine executed when connection is first established.

```
norm_str("Welcome to the WB2OSZ
        automatic frequency/
        deviation tester.");
norm_str(" ");
norm_str("Instructions:");
norm_str("  1. Send several
        non-blank lines.");
norm_str("  2. Wait for reply
        after each.");
norm_str("  3. Ignore any
        occassional erratic
        values.");
norm_str("  4. Disconnect when
        done.");
norm_str(" ");
norm_str("Recommended deviation
       is no more than 3 kHz.");
norm_str("Anyone who disagrees
        with the results is
        invited");
norm_str("to supply a reference
        more accurate than my
        2AT.");
norm_str(" ");
norm_str("               John");

/* Open the log file. */
/* Use the radio call with */
/* ".CAL" for file type. */

strcpy(temp,call);
strcat(temp,".CAL");

Fp_cal_log = fopen(temp,"a");

now = time(0);
my_time(&now,date_and_time);

fprintf(Fp_cal_log,
        "%s %s, %s\n", f_name,
        l_name, date_and_time);

/* Find out if direct */
/* connection or via */
/* digipeater(s). */

function_key (F_TNC_COMMAND);
hang_around (1);
function_key (F_CONNECT);
normal_key (CR);
hang_around (1);
function_key (F_CONVERS);

/* Another routine, that */
/* determines types of */
/* messages from TNC, stashes */
/* away the connection path. */
/* The closest digipeater, if */
/* any, is put in */
/* Digi_nearest and used in */
```

```
   strcat(s_compact, s_freq);
   strcat(s_compact, ", ");
   strcat(s_compact, s_dev);
   strcat(s_compact, "]");

   fprintf (Fp_cal_log,"%s\n",
              s_compact);

 /* Construct more self- */
 /* explanatory form for */
 /* report to user. */
 /* Instead of signed number */
 /* for offset, give absolute */
 /* value and "too low" or */
 /* "too high." */

   if (n_freq < 0)
    {
     strcpy(too, "low");
     for_mat(-n_freq, s_freq);
    }
   else
     strcpy(too, "high");

 /* See if direct connection. */
 /* If one or more digipeaters */
 /* used, put call of closest */
 /* one in the message. */

   if (Digi_nearest[0] == NUL)
     strcpy (who, "Your");
   else
    {
     strcpy (who, Digi_nearest);
     strcat (who, "'s");
    }

   sprintf(message, "%s frequency
         is about %s kHz too %s.
         Deviation is about %s
         kHz.", who, s_freq, too,
         s_dev);

   norm_str(message);
  }
```

fig. 5. Segment of a routine executed when a record is received from the station being tested.

for someone to connect. The user of the system is greeted with an explanation and a log file is opened (a program fragment is shown in **fig. 4**). For each record received, the A/D converters are sampled. The numbers are scaled to appropriate units, formatted into a message and sent back (a program fragment is shown in **fig. 5**). The user, date, time, and measurements are saved in a file for later analysis.

The method is certainly not foolproof. There is a chance that another packet's characteristics were sampled in the time that it took for the line of text to be

table 2. Newsletters dedicated to digital communications via Amateur Radio.

American Radio Relay League
225 Main Street
Newington, Connecticut 06111
Gateway: The ARRL Packet Radio Newsletter

British Amateur Radio
Teleprinter Group
7 Daubeney Close
Harlington,
Dunstable, Beds, LU5 6NF
DATACOM

Chicago Area Packet
Radio Association
21W464 Army Trail Rd.
Addison, Illinois 60101
CAPRA Beacon

Florida Amateur Digital
Communications Association
812 Childers Loop
Brandon, Florida 33511
FADCA>Beacon

Georgia Radio Amateur
Packet Enthusiast Society
P.O. Box 1354
Conyers, Georgia 30207
Grapevine

New England Packet
Radio Association
P.O. Box 15
Bedford, Massachusetts 01730
NEPRA PacketEar

Northwest Amateur Packet
Radio Association
13304 131st street KPN
Gig Harbor, Washington 98335
Zero Retries

Softnet User Group
Department of EE
Linkoping University
S-581 83 Linkoping, Sweden
Softnet News

Sydney Amateur Digital
Communications Group
P.O. Box 231
French's Forest, NSW
Australia 2086
The Australian Packeteer

Tucson Amateur Packet
Radio Corporation
P.O. Box 22888
Tucson, Arizona 85734
Packet Status Resister

Utah Packet Radio Association
4382 Cherryview Drive
West Valley City, Utah 84120
UPRA Connect

Vancouver Amateur Digital
Communications Group
9531 Odlin Road
Richmond, British Columbia,
Canada V6X 1E1
The Packet

transferred from the TNC to the computer. This is why instructions say to ignore any occasional erratic values.

One puzzled ham connected and got all kinds of random frequency reports, even though he hadn't sent anything. It turns out that a playful Amateur continually attempted to connect to my station while tuning his frequency back and forth. The solution was to recognize and ignore the "*** Connect request ..." message from the TNC.

There's also a manual mode that allows the system operator to perform measurements on another station by digipeating back to himself via the other station or just watching everything go by. The reporting and logging of measurements are naturally different, but the calculations are the same.

You don't need a high-powered computer for this application — something like a VIC-20 and a little BASIC program will be fine. It is important to get the line of text from the TNC quickly (at least 1200 baud) and sample the A/D converters before another packet comes along. After the data has been collected, there's plenty of time to do the calculations and prepare a response.

calibration

Proper adjustment is important to avoid giving incorrect reports. The procedure is simple. All you need is an accurate RF signal generator and voltmeter. I used the popular ICOM model 2AT synthesized RF signal generator as my source.

First set the signal generator (and receiver of course) to your local packet radio frequency. 145.01 MHz seems to be the most popular frequency in most parts of the country. Adjust R6 for zero volts at pin 7 of U1. Set the signal generator frequency 5 kHz higher and adjust R3 for 2.5 volts at the same place. Finally, set the frequency 10 kHz lower (i.e., 5 lower than original) and observe the voltage. It should be close to −2.5 volts.

If the negative voltage is much different than −2.5, you have a linearity problem as I did. In this case, alter the setting of R3 until you set a DIFFERENCE of 5 volts between + and −5 kHz input. At least this will give a fairly accurate deviation for signals on frequency.

improvements

The output from my receiver is not very linear. For instance, a station 2 kHz too high might be given a report that it's 2.5 kHz too high, while a station 2 kHz too low might be told he's 1.5 kHz off frequency. (Note: a person on frequency is told he is on frequency. The problem is non-linearity, not an offset of 0.5 kHz.) Possible solutions are compensation in software, repair of the radio — or purchase of a new radio.

The modems commonly used for packet require the two tones to be fairly close in amplitude. Hank, WØRLI, suggested measuring the amplitudes of the

tones separately. This would require two band-pass filters (for 1200 and 2200 Hz) instead of the high-pass filter made up of C3 and R12. An additional peak detector, low-pass filter, and A/D converter would also be required. The deviation measurement would be based on the larger value. The difference in amplitude could be reported something like, "Amplitude of 2200 Hz tone is 89 percent of other tone."

conclusion

Asking people to drag their equipment to a meeting for adjustment was not successful; the automated approach has produced much better results. During the first few days of operation about 20 stations tried out the system. (This might not sound like many but I'm running only 1-1/4 watts in a valley.) During later measurements, most stations that were substantially off had made correcting adjustments.

For readers who wish to become better informed about packet radio, a list of organizations with newsletters oriented toward digital communications via Amateur Radio is shown in **table 2**.

acknowledgements

Thanks to Gary, WA1GRC, for the idea and for nagging us at every NEPRA meeting until someone actually designed and built the unit.

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ARRL, *Second ARRL Amateur Radio Computer Networking Conference, March, 1983*, American Radio Relay League, Newington, Connecticut, 1983, ($9.00).

ARRL, *Third ARRL Amateur Radio Computer Networking Conference, April, 1984*, American Radio Relay League, Newington, Connecticut, 1984 ($10.00).

ARRL, *Fourth ARRL Amateur Radio Computer Networking Conference, March, 1985*, American Radio Relay League, Newington, Connecticut, 1985, ($10.00).

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

low frequency DX

No more about Ye DX bands
Do Wilde Men push and pulle,
Or talk about TA's, UJ's,
Or toss about Ye Bulle.
Ye Sunspot Cycle hath gummed Ye Game,
Ye bands are dry as Snuffe.
And many Hardy Souls, no doubt,
Will find it hard to do Without,
Excepting Thee and Me, Old Friend,
Who Never Worked Ye Stuffe.

This touching ode to DX, written by By Goodman, W1DX, in the late 1940s, certainly applies to DX today, as far as I'm concerned.

But while HF DXers may be moaning and groaning over the medicore conditions, low frequency DXers are having a fine time. Eighty and 160 meters are *jumping* these days!

A note from Bob Eldridge, VE7BS, tells of some of the DX worked in the Pacific Northwest:

The Australian CW signals can be anywhere, usually below 1825 kHz. The VK SSB rag-chewing groups on daily are most often about 1832 kHz and 1825 kHz, with a few around 1815 kHz. They are most consistent this time of year (early summer) when they can be worked virtually every morning (their evening). I hear them from about half an hour before sunrise to about half an hour after. At about the time they come up in strength here they are dropping out at WØZV (Colorado), and he drops out to me at about the same time. He also hears them before and after his sunrise, so the VK opening to North America on 160 meters is quite long.

But although the Spring equinox is the most reliable time for Pacific contacts, I notice looking back through the log that I had good DX contacts in January and February and some DX in every month of the year. This week (week of July 24th) everyone is on the lookout for T31AT (Kiribati) and A35PP (Tonga).

VE7BS uses an inverted-V antenna with the apex at 105 feet and the ends about 60 feet (**fig. 1**). This is his "comparison antenna" for the others that he has experimented with from time to time. Bob says it is broadside to Australia and Europe, but works reasonably well in all directions. Three parallel wires are used in each leg and the coax feedline is wound into an RF choke just below the feedpoint of the antenna.

Another 160-meter antenna that VE7BS has used with success is the so-called "Lazy-U," shown in **fig. 2**. The vertical portion can be from 50 to 100 feet long, with the horizontal portions bringing the system to resonance without contributing much horizontally polarized radiation. The VE7BS "Lazy-U" worked better than the inverted-V in some directions and this was the antenna he used to land 5N8ARY (Nigeria).

An interesting adaptation of this vertical antenna that some 160-meter DXers use is the so-called "G8ON" antenna, named after the Amateur who popularized it on the band (**fig. 3**). The antenna is a half-wavelength long, with the high current portion in the vertical plane. The wire is end-fed from the top end.

In closing, Bob has some interesting remarks about radial systems, as applied to 160-meter antennas. He advised the 160-meter operator not to worry too much about extensive radial systems. He says:

I managed WAC on 160 meters with a vertical without any radials and K7VIC has one of the most potent signals on the band using a vertical top-loaded monopole without radials, so I wouldn't get depressed if I had no room for radials. For transmitting, I see nothing wrong with a 45-foot tower, top-loaded with a Yagi, working against a few properly disposed 8-foot ground rods near the base and a cluster of short radials or chicken wire mesh under the tower. As far as I can see, the main disadvantage of a relatively short, loaded vertical is the narrow bandwidth achieved without retuning.

For those west coast DXers interested in 160 meters, Bob recommends (and I concur) the *160 meter West Coast Bulletin*, published by Dennis Peterson, N7CKD, 4248 A Street SE, Space 609, Auburn, Washington 98002. An SASE might bring you details from N7CKD.

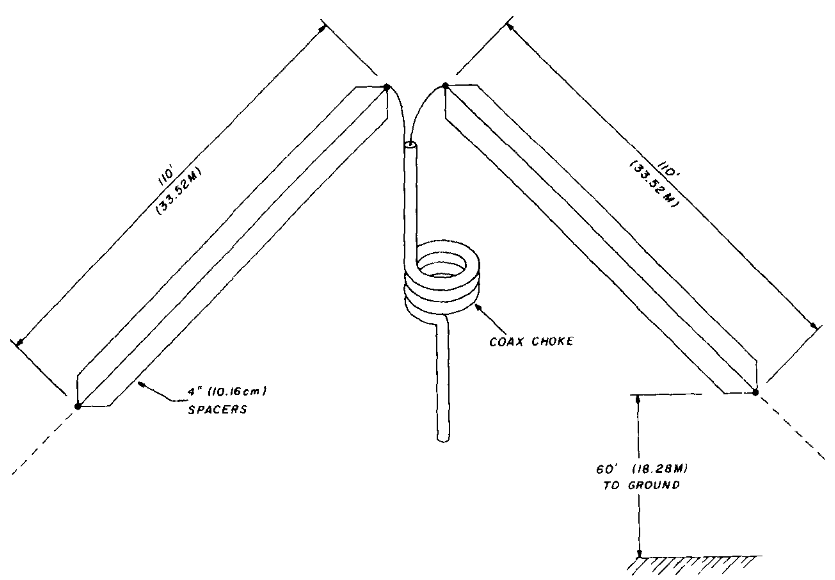

fig. 1. The inverted-V antenna for 160 meters at VE7BS. Wires all lie in the same plane. Apex is at 105 feet (32 meters).

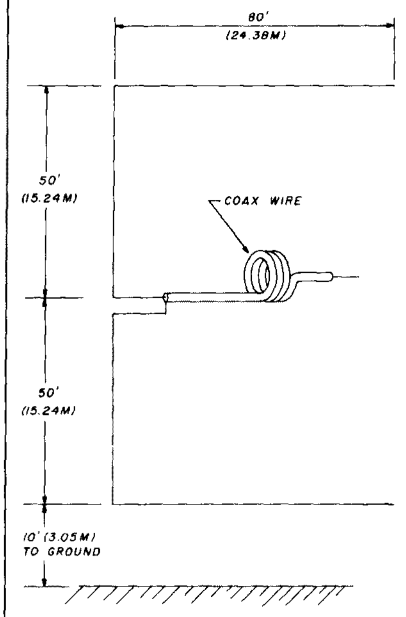

fig. 2. The VE7BS "Lazy-U" antenna for 160 meters.

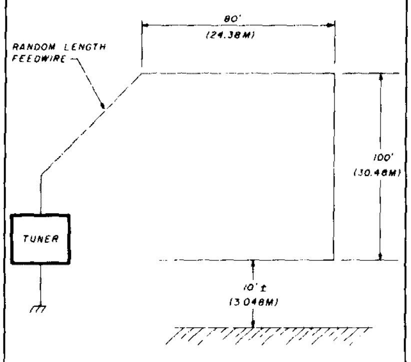

fig. 3. The G8ON vertically polarized antenna. The vertical section can be as short as 33 feet (10.5 meters.)

the effects of trees and vegetation on your signal

From time to time I've received inquiries from Amateurs asking what effect upon their signals a nearby tree, or group of trees, might have. Since I didn't know, I could only reply with an evasive, ambiguous answer. My good friend Marv, W6FR, who was "bugged" by a tall tree into which his 20-meter beam fired on the European path, was convinced that during the months the tree was in bloom with heavy foliage, his signal suffered. When pressed for specifics, however, he admitted under pressure that his often-stated conclusion was a hunch. The upshot of this was that his local DX competitors felt they had a psychological edge on Marv during the spring months when the tree was in its full glory!

The June/July issue of *Broadcasters ID* (published by Information Dissemination, 2501 Hilldale Boulevard, Arlington, Texas 76016) has some interesting information on this subject. In an article by E.J. Pryor, Jr., of Broadcast Technologies, Inc., the subject of foilage and vegetation is discussed, with respect to AM and FM broadcasting. The article says, in part:

Almost any engineer at an AM directional station could tell you that his array shifts each year due to the many factors which are related to seasons.

In most cases, some of the changes can be traced to ground conductivity which varies due to the moisture, water table variance, and the temperature factor in the area. Foliage growth has a direct effect on the radiation performance of your transmitting antenna....

The vegetation surrounding your transmitter plant absorbs and reradiates some of the energy radiated by your antenna. At AM frequencies, the vertical field can be reduced significantly by high grass and green trees near the antenna farm. At FM frequencies, this signal loss can be approximately 2.5 dB. Above 1000 MHz, the losses due to ground scatter, signal absorption, Fresnel zone losses and terrain can drastically change with the green season. Losses can be as much as 10 dB, or more.

You can control the foliage on your property where your transmitter and tower are located. Regardless of your frequency, the area around your transmitter plant should be mowed regularly and kept free of trees. Trees have the greatest effect on your signal.... Tall grass, especially when green or freshly wet, can detune a directional array, upset drive point impedance, mutual coupling factors and significantly degrade the station's performance.

While these remarks are aimed at

vertical AM broadcast antennas and FM arrays, the ideas could apply to Amateur antennas. Most Amateur HF antennas are horizontally polarized, and my personal opinion (apologies to W6FR) is that nearby trees and foliage have relatively little effect on antenna performance in the HF region. In the case of vertical antennas, however, Mr. Pryor's remarks my be interpreted to mean that foliage and tall grass can affect the operation of the vertical antenna in the HF/VHF spectrum as well as in the broadcast band.

I would appreciate hearing from readers who may have experience in this area to find out what effect, if any, nearby trees and bushes have on the operation of both horizontally and vertically polarized Amateur antennas in the HF and VHF regions.

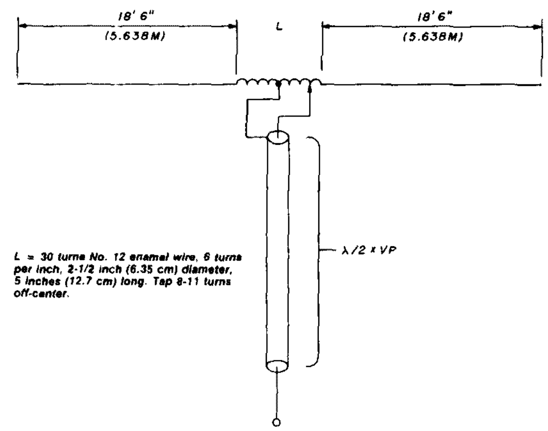

fig. 4. The "shorty-forty" dipole of WØSVM.

the WØSVM "shorty-forty" dipole

A good idea and a catchy name! Jack, WØSVM, has spent considerable time and effort designing a compact, practical dipole antenna for city dwellers who don't have the space to put up a full-size 40-meter antenna and for various reasons don't want to use a vertical antenna.

Jack wanted to build a simple, rugged antenna that would not have loading coils flopping around out in the elements. He felt that a center loading coil could do the job, if the coil was made properly. After several months of experimentation, he came up with the antenna shown in **fig. 4**. Briefly, it's a center-loded dipole about half the size of the full dipole. The feedline is tapped on the loading coil in such a manner as to provide a good match to a 50-ohm line. The line is cut to an electrical half or full-wavelength and can be run directly to the transmitter or to a transmatch for maximum frequency flexibility. When used with a simple transmatch, the antenna has a 500-kHz passband between the 1.5:1 SWR points.

The resonant frequency of the antenna is determined by the wire sections. The difference in tip lengths between 7.0 and 7.3 MHz is 10 inches. It's best, therefore, to cut the antenna to that portion of the band in which most of the operating is to be done. Without the transmatch, bandwidth of the antenna is about 150 kHz between the 1.5:1 SWR points.

The shield of the coax line is tapped to the center point of the coil and the center conductor is tapped off-center. Using the 18-foot, 6-inch (5.638 meters) flat-top dimensions, the coil is tapped 11 turns off-center for operation at the low end of the band and at nine turns off-center for operation at the high end of the band.

Exact antenna resonance and the minimum value of SWR can be achieved at any point in the band by changing the tip length of the wires and the feedpoint on the coil. If operation is mainly confined to the high frequency end of the band, the wire sections can be reduced in length to 17 feet 8 inches (5.384 meters). All in all, the design is quite flexible and the resonant frequency and impedance match can be varied at will to suit any spot in the band, and also to match a 75-ohm transmission line, if desired.

The antenna can be erected in the conventional fashion or made into an inverted-V, with the end tips close to ground level. For best results, the center of the antenna should be from 30 to 50 feet (9 to 15 meters) in the air, and relatively clear of nearby objects.

While WØSVM doesn't mention it, I've found it helpful in some cases to wind the feedline into a choke coil directly below the antenna feedpoint. This reduces the RF field on the outside of the coax line and can reduce TVI in some instances. Of course, if the feedline is run parallel to the antenna after the choke is installed, all bets are off because the antenna will be coupled to the feedline by mere proximity. It's best to bring the feedline down vertically below the center of the antenna to ground level, or to the level of the station, if it's located on a higher floor. Running the feedline parallel to the antenna element(s) is bad practice, regardless of the type of antenna used.

do you have an unusual antenna?

Do you have an unusual antenna installation that would be of interest to readers? If so, I'd like to see it. Just send a clear pencil sketch of the antenna, including dimensions and the electrical characteristics, such as the SWR or operating bandwidth. A good black-and-white photograph is always appreciated, if the antenna can be photographed! (It's very difficult to take a decent picture of a wire antenna — although I have a friend who got

good pictures of a wire antenna by taking the picture at night, using a camera with a flash attachment.)

Antennas featured in this column will win their owners an autographed copy of my *Beam Antenna Handbook.** For those who don't have an inspirational antenna to talk about, the handbook is available from Ham Radio's Bookstore.

MXHNY

In closing this December column, I wish my readers a Merry Christmas and a Happy New Year. And may DX be good to you in 1986!

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*Available from Ham Radio Bookstore, Greenville, New Hampshire 03048, $9.95 plus $3.50 shipping and handling.

ham radio

Determining the next step in flexible, robust digital communications

AMTOR, AX.25, and HERMES: a performance analysis of three systems

AMTOR and the AX.25 packet protocol are currently being heralded as the state-of-the-art in Amateur digital communications. This article reports the results of an objective performance analysis of these two systems, and compares each to HERMES, a third, newly proposed system. Four performance measures are described and applied to each system.

applications

In order to choose reasonable conditions under which to evaluate the performance of competing communications systems, it is first necessary to look at the ways Amateurs actually use digital communications.

If you listen to normal RTTY traffic on the HF bands, you'll find that most Amateurs are engaged in casual conversation. These QSOs are almost exclusively conducted at a 45 baud (60 WPM Baudot) channel rate, with throughput usually limited by the speed at which the respective operators can type. While some of us can type fast enough to keep a 45 baud system running fairly continuously, most cannot.

Other users — though fewer in number — are involved in RTTY traffic nets, using computers or RTTY equipment to relay third-party message traffic. Although surprisingly little RTTY traffic handling actually occurs (compared to the amount handled by CW or SSB) at this time, the availability of improved digital communications schemes may help to encourage the growth of this type of activity.

Some users employ computer data transfer for traffic handling. It is primarily this application for which the AX.25 "packet" systems are designed. And on VHF, packet radio activity is growing rapidly. AX.25 has the capability of supporting conversation just like RTTY, as well as direct computer-to-computer data file transfer.

I believe we should push for a single digital communications scheme that can adequately support all types of users at both HF and VHF, including satellite links. Such a system is not currently in use, nor has one yet been proposed.

Two of the three classes of users described above are dealing exclusively with plain language text, using the Baudot alphabet (or the AMTOR variation of the Baudot alphabet). Computer hobbyists are using ASCII, sending it in an eight-bit format so that arbitrary computer data (as well as text) can be transmitted within that scheme as well. Whatever digital scheme we agree to accept as standard, it should support both types of alphabet.

Those using digital communications for casual conversations are probably not too worried about a few errors now and then in the received text, but would like to see the channel processing their data fast enough (100 bps is probably sufficient). On the other hand, the other two user classes have no tolerance for any hits, and are willing to sacrifice some channel throughput to attain the required reliability. It would be best to have a system that could be conveniently optimized by the operator for each of these different applications.

comparing systems

The following four criteria are useful in comparing the various systems available:

• **Throughput under ideal channel conditions.** An "ideal" channel introduces no errors, thereby allow-

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ing the communications system to run at its highest possible rate. (This is expressed in characters per second (cps).)

• **Robustness.** This is the probability, expressed as a percentage, that the system will falsely accept random noise or badly corrupted characters as valid. Any system must reject (or correct) all corrupted data received under these conditions with high reliability — i.e., have a very low probability of falsely accepting corrupted data characters as valid. (The mathematics used to compute this probability for each of these two systems evaluated is presented in the appendix.)

• **Bit Error Rate (BER) required to stop progress (expressed as a percentage).** It is assumed that random bit errors are occurring at a certain rate. How high must this error rate be to prevent the system from occasionally transferring data successfully? Under these circumstances, the throughput of the communications system is sharply reduced since there are many repeats. But we want the system to make *some* progress — occasionally, data should be correctly transferred and acknowledged by the receiving end. Other things being equal, it is desirable for a system to be able to tolerate as high a random BER as possible before forward progress is stopped.

• **Minimum Required Error Free Seconds (MREFS)** to maintain progress in forward data transfer. Here a channel generating "burst" errors is assumed. The channel makes no errors for a certain length of time, and then becomes unusable for a certain length of time (an error burst). An interesting parameter of a communications system is how short the "good" period of the channel may be while still allowing successful and correct data transfer to occur occasionally. Other things being equal, a system should have a low MREFS requirement.

To make reasonable comparisons, we will assume that each of the packet protocol systems is being operated with the same type of channel, arbitrarily a 100 bps synchronous channel such as that specified for AMTOR, with a 20 millisecond allowance for radio turn-around from transmit to receive and vice-versa. For consistency, it is assumed that a 5-bit symbol alphabet (Baudot) is being used when specifying throughput performance, even though any of the systems can transfer Baudot, ASCII (7-bit), or arbitrary computer data (8-bit) by employing appropriate alphabet conversion subroutines.

Figure 1 illustrates AMTOR in operation.[1,2] In the **ARQ** mode, AMTOR sends three 7-bit characters and then pauses for an acknowledgement signal (one 7-bit character) from the receiving station before proceeding with the next three-character group. If no acknowledgement character is received, the last three-character group is repeated.

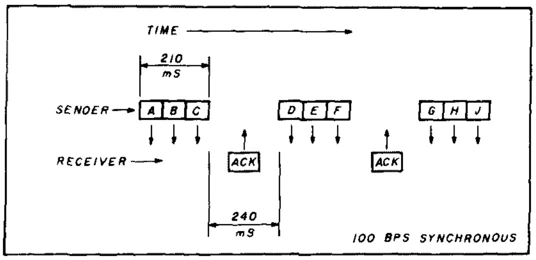

fig. 1. AMTOR ARQ mode.

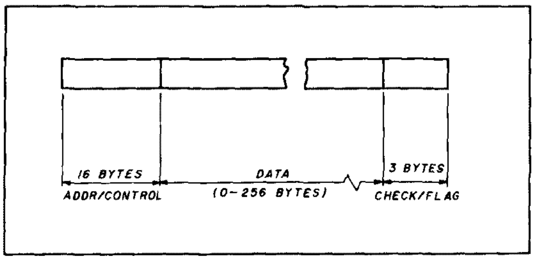

fig. 2. AX.25 packet format (no repeater ID bytes).

Time relationships are quite specific. AMTOR spends 210 milliseconds sending each group, and 240 milliseconds waiting for the acknowledgement before sending the next group.

Like Baudot, AMTOR uses a 32-character alphabet, but uses 7-bit symbols instead of the Baudot's 5-bit symbols. Each AMTOR symbol is composed of three 0 bits and four 1's, allowing it to detect errors in its received data (sometimes). To do this, AMTOR uses a simple parity check. Each received character must have four 1 bits and three 0 bits; if it doesn't, an error is flagged, and the entire group is discarded and must be repeated.

If the channel makes one bit error per character (or any *odd* number of bit errors per character) this simple parity check successfully detects the error. But if the channel reverses two (or any *even* number) of the character's bits, its 4/3 parity ratio will be preserved even though it has now been transformed into a different character of the alphabet. Under these conditions, the AMTOR code will fail to detect the error.

With the channel producing few errors, it is likely that no more than 1 bit error per character will be experienced. When the channel is very poor, however, and is making many bit errors, one has about an equal chance of experiencing an odd or even number of errors. Consequently, there is a 50 percent chance of an incorrect character unintentionally satisfying the parity check.

packet system design

Good references for a detailed description of the AX.25 or "packet protocol" are available.[2,3] AX.25 is an adaptation of the 15-year-old data communicatons protocol pioneered by the Defense Advanced Research Projects Agency in the 1960s for error-free communications within telephone computer networks. This protocol assumes a telephone channel, or a channel of similar quality (in terms of signal-to-noise ratio, bandwidth, and lack of interference) in its design.

Figure 2 shows the current AX.25 packet makeup: 16 bytes or more of synchronizing and header information at the beginning of each packet, followed by up to 256 bytes of information, with the packet finished out with three final bytes for error checking and flagging the end of the frame. If the packet has been processed by one or more repeaters, seven additional address bytes are added to the packet to identify each repeater, up to a maximum of 9. For purposes of this discussion, use of a simplex channel only, with its requirement for only 19 overhead bytes in an AX.25 frame, is assumed.

Although AX.25 includes a significant amount of overhead in each packet, it is a very good system in light of the environment for which it was originally designed. It is not, however, ideal for use on a channel that doesn't look much like a telephone channel — a narrowband HF channel with fading, noise, and a high error rate, for instance.

Because this system was designed for transmitting computer data, it contains a robust error detecting scheme that provides a good probability of detecting a garbled block regardless of the source or severity of the errors.

Figure 3 shows the flow of activity on a packet channel. Since just one block at a time is sent, and each block must be acknowledged by the receiving station before the next block is sent, a certain amount of channel time is inevitably spent waiting. For our analysis, delays similar to those used with AMTOR are assumed.

The packet protocol is somewhat adaptive. Depending upon channel conditions, the operators can adjust the length of each packet by controlling the number of data characters sent (up to 256) in each block. Thus, when conditions are good and the channel is rarely making errors, a full-size block may be used, with a correspondingly small proportion of the channel time wasted in packet addressing overhead and waiting for acknowledgements. On the other hand, when the channel is very bad, the character count can be greatly reduced to shorten the packets, thus improving the chance of their being received error-free. *Very* short packets are quite inefficient, however, because of the addressing bytes that must always be included.

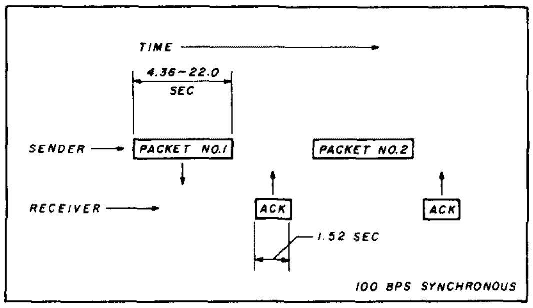

fig. 3. AX.25 channel activity.

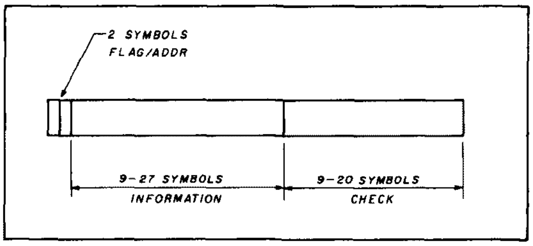

fig. 4. HERMES frame format (5-bit symbols).

For our purposes, then, we will examine the packet systems running with both maximum length packets (even though, in practice, hams rarely use more than about 80 data characters per packet) and extremely short packets (only 16 data characters) as well.

HERMES system design summary

HERMES was designed to be superior to both AMTOR and AX.25 in the Amateur narrowband HF environment. The latest version of a system described in reference 4, HERMES has been used experimentally for the past several years.

The key to HERMES is the use of a Reed-Solomon forward error-correcting code. The chosen code uses a 5-bit symbol alphabet and 31 symbols per block; a typical frame is diagrammed in **fig. 4**. The coder and decoder are *adaptive* — i.e. the ratio of data symbols to check symbols in each block is controllable by the operators to allow optimization to channel conditions. The check symbols add redundancy to the data in a special way that allows the system to mathematically correct some symbols in each frame that have been altered or destroyed by the channel. (Reference 4 also provides additional details on how forward error correcting systems work.)

This feature allows HERMES to perform efficiently when channel conditions are favorable, as well as

when the channel conditions are poor (in this case data transfer efficiency is traded off in favor of gaining additional error correcting capability).

HERMES uses a protocol that assumes a single link (i.e., a pair of stations in contact), but is also designed to accommodate net type operations. Unlike AX.25, it is not configured to allow several simultaneous and independent QSOs on a single channel, thus greatly reducing the number of overhead bytes that must be transmitted in each block.

The system is designed to handle Baudot characters, and conversions of 7-bit ASCII characters and arbitrary 8-bit bytes directly, so that all types of data can be handled with the same efficiency.

The operator can choose any of 48 different configurations that have been implemented; these support 16 different modes in each of the Baudot, ASCII, and byte communications modes. Data transfer efficiency of the system ranges from 93.5 down to 32.3 percent, depending on the degree of error correcting capability chosen. The probabilities of falsely accepting corrupted data as valid vary from 2.6 to 0.0000009 percent; error correcting capabilities vary from 0 (error detection only) to 11 symbol errors per block (35 percent correction capability).

The system can function in an **ARQ** mode or in a broadcast mode in which no acknowledgements are sent. In the **ARQ** mode, HERMES sends from nine to 27 data blocks per transmission (depending upon system configuration), after which the link is turned around, and the receiving station sends one "acknowledge" frame to identify all the blocks that were received correctly or were correctable. This pattern is illustrated in **fig. 5**. Since the ratio of data frames to acknowledge frames is quite high, there are relatively few link turn arounds and little wasted link time spent waiting for radios to switch. In the broadcast mode, HERMES sends the data frames using the same code configurations, but does not wait for acknowledgements.

AMTOR (ARQ) analysis

AMTOR sends three characters (requiring 210 milliseconds), then pauses for 240 milliseconds to allow the acknowledge signal to be received. Therefore 3 characters require 450 milliseconds to send under ideal conditions, resulting in a 6.67 cps throughput.

Under conditions of no signal (random noise input only), AMTOR will recognize a "block" only if it sees three valid characters. Since there are 34 legal characters out of the 128 *possible* 7-bit characters, the probability of any single character looking valid, with random input, is 34/128 = 0.266. The probability that three such characters in a row are received with only noise input is the third power of this number, or about 1.9 percent.

Under very noisy signal conditions, in which we assume that many bit errors are being experienced, and we assume that a valid signal is being received in addition to the noise, we have a 50 percent chance of experiencing an even number of bit reversals in the received characters. If the odds of receiving an incorrect, but valid-looking, character are 50 percent, then the probability of receiving three of these is the third power of 0.50, or 12.5 percent. This second case is really more relevant to our discussion, and it is this number that we'll use for our robustness figure for AMTOR.

Just one bit error during each 450 millisecond is enough to stop data transfer progress with this system. This bit error would either corrupt the data transmission or the acknowledge signal, and the corruption of either is sufficient to force a repeat. Once we continually force repeats, the forward transfer of data has stopped. One bit in 450 milliseconds at a 100-bps data rate corresponds to a 2.22 percent BER.

To get data through once in a while, AMTOR needs to occasionally obtain a 450 millisecond window of *error-free transmission* by the channel. Therefore, for this system, MREFS = 0.45.

maximal frame packet analysis

A maximal length frame (256 data bytes) is used to configure the packet system for performance under ideal conditions. Without the repeater addressing bytes, the standard AX.25 packet requires 19 bytes of overhead for addressing and error checking, so the complete frame is 275 bytes long.

The AX.25 acknowledge frame would consist of just the 19 overhead bytes. With 22.0 seconds required to send the 275 byte information packet at 100 bps and 1.52 seconds required to send the acknowledge packet (plus two 20 millisecond intervals assumed for radio switching), the system requires 23.56 seconds to transfer one frame under ideal conditions. This corresponds to 10.86 cps for 8-bit characters, or 17.36 cps for equivalent 5-bit characters (409 5-bit characters can be loaded into the 256 byte data frame).

By the nature of the 16-bit CRC (Cyclic Redundancy Check) code used for error detection in the AX.25 format, the probability of a corrupted block being falsely accepted as valid is 1.53E-5 (0.00153 percent), regardless of whether we are talking about random noise or a noisy signal input to the system. The CRC is a much more sophisticated algorithm than the simple parity check used in AMTOR, and is much more robust in the presence of massive channel errors.

Since AX.25 is only an error *detection* scheme, *one* bit error occurring during every 23.56 seconds would destroy the correctness of either the data frame or the ACK frame, and in either case the data frame would need to be repeated. Therefore, a random BER of

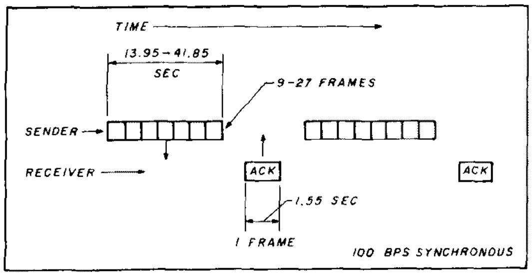

fig. 5. HERMES channel activity.

1/2356 or 0.042 percent would stop all progress for AX.25 in this configuration.

By the same token, to get data through once in a while, the channel must occasionally be good for at least 23.56 seconds at a time. So MREFS = 23.56.

minimal frame packet analysis

When channel conditions are very poor, we would want to operate the packet system with a very short frame — let's say just 16 data bytes, in addition to the 19 overhead bytes always required.

With this frame length, we need only 2.8 seconds to send the data frame, and a total of 4.36 seconds to complete a data transfer, including the ACK frame. This corresponds to 3.67 cps for 8-bit data and 5.87 cps for 5-bit characters.

The probability of a corrupted block being falsely accepted as valid is the same as with the maximal frame configuration, 0.0015 percent since this characteristic depends only upon the number of check bits used and the error detection algorithm.

The BER required to stop the system is now 1 bit in 4.36 seconds, or 0.23 percent. MREFS for this configuration is 4.36.

HERMES analysis

Case 1. For the first HERMES configuration we assume a good quality channel and choose a mode appropriate for maximum data throughput and adequate error probability for conversational use. We use a configuration with 29 data symbols and just 2 check symbols per block and operate the decoder in an error-detecting mode only.

For the throughput calculation, we assume the transmission of 27 data frames containing 27 information characters each, followed by 1 acknowledge frame from the receiving station. After allowing for two 20 millisecond switching intervals for the radios, the total time required for the exchange is 43.44 seconds. Since 729 characters are transferred, throughput is 16.78 cps (5-bit characters).

With two 5-bit check symbols and an error detecting mode only, the probability of falsely accepting a bad block is 0.098 percent, regardless of whether we are talking about random noise or a noisy signal being fed in to the system.

To stop forward data transfer, we must have 1 or more bad bits in each frame sent by the sending station, which corresponds to one bit error out of 155, or a 0.65 percent BER. We can make some progress as long as at least one frame can get through once in a while, which requires 1.55 seconds, so MREFS = 1.55.

Case 2. For our second example we choose a HERMES configuration with 17 data symbols and 14 check symbols in each frame, with the decoder running in a six-error correcting mode (6 symbols out of each 31 symbol frame can be corrected).

In this configuration, the system will send up to 15 data frames before waiting for an acknowledge frame from the receiving station. This works out to 24.84 seconds to transmit the 15 frames and receive the acknowledgement. 225 characters would be transmitted during this time, for a throughput under ideal conditions of 9.06 cps.

In this mode, the decoder's probability of falsely accepting a bad block as valid is 0.000059 percent.

Since six symbol errors can be made in each frame with the decoder still being able to correct the block, this corresponds to one symbol in every five being in error. In order to stop forward progress, then, we must have one character in every *4* be in error. This corresponds to a random bit error rate of 1 in 20, or 5 percent.

On a bursty channel, we must be able to receive at least 25 symbols of a block without error in order to be able to correct it completely. This corresponds to MREFS = 1.25 seconds.

Case 3. For this example we assume a very poor channel, and are willing to sacrifice additional throughput to enhance the forward error-correcting power of the code. Here we use a configuration with 11 data symbols and 20 check symbols per block, and we allow the decoder to correct up to 10 errors per frame.

In this configuration, HERMES will send 9 data frames at a time before waiting for an acknowledgement, and spend 15.54 seconds doing it. In these 15.54 seconds, 81 characters will be transferred, for a throughput of 5.21 cps.

In this configuration, the probability of the decoder falsely accepting an invalid block is 0.0000029 percent.

Since the decoder can correct 10 symbols out of a 31 symbol frame, a channel making random bit errors can destroy every third symbol, and the decoder will still be able to fully correct the frame. Therefore, to stop the system, the channel must corrupt one symbol out of every two, for a random bit-error rate of 10 percent.

table 1. Performance summary.

system/statistic	throughput (cps)	robustness (percent error)	BER required to stop (percent)	MREFS (seconds)
AMTOR ARQ	6.67	12.5	2.220	0.45
AX.25 (Max)	17.36	0.0015	0.042	23.56
AX.25 (Min)	5.87	0.0015	0.230	4.36
HERMES/1	16.78	0.098	0.650	1.55
HERMES/2	9.06	0.000059	5.000	1.25
HERMES/3	5.21	0.0000029	10.000	1.05

On a bursty channel, we need to get 21 symbols out of every frame transmitted without error, so MREFS = 1.05.

performance summary

Table 1 summarizes the performance figures we have developed for each of the competing schemes. Once again, the ideal system would have a high throughput, a very low robustness percentage, a very low BER required to stop, and a very low MREFS.

AMTOR can absorb up to a 2.22 percent random channel BER before being stopped, and needs only 0.45 seconds to make progress, which is good, but the fact that it has a 12.5 percent chance of falsely accepting invalid data as valid is disqualifying. We can, and should, do much better than that.

The two AX.25 packet configurations evaluated, which fully bracket the usual operating configurations, represent the extremes of performance available with the AX.25 protocol. First, we see that packet's probability of falsely accepting invalid data is fairly low, which is good. 0.0015 percent is low enough for most purposes, and might need augmentation only when very large files are transferred at high data rates at UHF and beyond. (This figure equates to the acceptance of one bad frame in about 67,000.)

The maximal packet configuration produces a good throughput figure of 17.36 cps, but at the expense of allowing only a 0.042 percent random BER before being stopped and requiring 23.56 seconds to get a packet through. As we said before, if you have a good enough channel, this will work nicely. Good channels are pretty easy to get at VHF, but HF is another story.

The minimal packet configuration allows the channel BER to increase by a factor of 5, up to 0.23 percent and reduces the minimum required error-free seconds to 4.36, which is more realistic for an HF channel. The throughput, however, has now dropped to 5.87 cps.

HERMES configuration 1 was chosen to obtain the highest possible throughput while maintaining a robustness adequate for conversational communications. It achieves a 16.78 cps throughput, which is very nearly as good as the AX.25 protocol under the best of conditions, and it does this while allowing a 0.65 percent BER before being stopped (this is 15 times more tolerant than the maximal length packet scheme), and an MREFS figure of just 1.55 seconds (1/15th the minimum required time for maximal length packet). Robustness in this configuration, 0.098 percent, is not as good as the AX.25 figure, but is more than adequate for the intended application, casual conversation.

Comparing the performance of HERMES case 1 to the minimal length packet case, we see that the packet scheme runs at about one-third the throughput and is still about three times less tolerant in both BER and MREFS.

HERMES configuration 2 was chosen for use on a moderately degraded channel, and in an application where high accuracy was required. Its 9.06 cps throughput falls midway between the extremes of the AX.25 configurations and its robustness is several orders of magnitude better than AX.25. Interestingly enough, it does this while being even more tolerant of channel errors, now allowing a 5 percent random BER before being stopped, and requiring 1.25 error-free seconds to transfer data. Comparing these figures to those for the minimal length packet scheme, HERMES is providing 54 percent more throughput while allowing a 27 times greater random BER in the channel, and requiring "good channel" bursts only one-third as long as AX.25. All these factors translate to superior performance by HERMES.

Even though the performance of AMTOR is disqualifying, due to its poor robustness, it is interesting to note that in this case, HERMES provides 36 percent more throughput. And while AMTOR does excel in the MREFS department, requiring only 0.45 seconds compared to 1.25 for HERMES, HERMES will allow more than twice as high a random BER before being stopped (5 percent versus 2.2 percent, overall), therefore, HERMES wins the comparison here, too.

Chosen for use on a very bad channel, HERMES case 3 allows a throughput of 5.21 cps, slightly worse than the minimal length packet case. But it has a phenomenally low probability of falsely accepting invalid data (0.0000029 percent), and can withstand a

10 percent random BER, 50 times that tolerated by the AX.25 scheme even with minimal length packets.

conclusions

Although AMTOR has excellent burst error performance, its probability of falsely accepting invalid data under poor conditions (at 12.5 percent) makes it a non-competitor when systems like AX.25 and HERMES are considered.

Under good channel conditions the AX.25 packet scheme does very nicely, with good throughput and adequate robustness, but it bogs down rather rapidly once channel conditions start to degrade, since it has no capability for forward error correction.

HERMES combines, in one adaptive system, the capability to achieve very nearly the same throughput as AX.25 under ideal conditions, as well as very nearly the burst error performance of AMTOR. It allows convenient optimization and is able to tolerate a much higher rate of random channel errors than is either AMTOR or AX.25 due to HERMES's use of powerful Reed-Solomon forward error correcting codes. It represents the next step in flexible and robust digital communications.

Appendix

The probability of falsely accepting a corrupted data block (an "Undetected Bad Block") will be denoted as P_{UBB}. It can be computed as follows for an error-correcting or error-detecting algebraic block code (such as the CRC or Reed-Solomon codes.)

For an *error-correcting* code:

$$P_{UBB} = \frac{\binom{A-1}{E} \cdot (A-1)^E}{A^C}$$

where: A = 2^m (the code alphabet size)

m = the number of bits per codeword symbol (m = 8 for AX.25 CRC, m = 5 for HERMES)

$A-1$ = $2^m - 1$

E = the number of errors being corrected by the decoder

C = the number of check symbols in each codeword

$\binom{A-1}{E}$ = the combinatorial factor for (A-1) things taken E at a time $= \frac{(A-1)!}{E!\,(A-1-E)!}$

For an *error-detecting* code:

$$P_{UBB} = 1/A^C$$

where: A and C are defined as above for error-correcting codes.

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ham radio

PRACTICALLY SPEAKING ... Joe Carr K4IPV

tracking the hideous intermittent — part 2: thermal intermittents

In Part 1 of this two-part series we dealt with the problem of troubleshooting mechanical intermittents. In Part 2 we'll discuss thermal intermittents and their solution.

Both heat and cold can affect a piece of electronic equipment for the worse; in its most blatant form, the set refuses to operate properly under either hot *or* cold conditions. During the winter Amateur mobile equipment is subjected to the *local* overnight temperature, which may be as low as 40 degrees below zero in some areas. During the summer, on the other hand, mobile equipment will be subjected to temperatures considerably above local air temperatures. In 1963, when a major automobile electronics company began experiencing reliability problems with its new solid-state models, it asked employees to leave their cars unlocked so that the engineers could measure the cabin temperatures. After four hours in 90-degree sunlight, the interior temperatures were found to be 140 degrees at the front seat and up to 180 degrees behind the dashboard!

These extremes of temperature can result in some peculiar intermittents. One familiar form is the set that won't work when you get into the car, but will work ten minutes after the heater or air conditioner has altered the cabin temperature.

Equipment used at the home or base station doesn't suffer the extremes of ambient temperature, but nonetheless may experience temperature-related intermittents. Typically, a set will either fail to work at all until *it* heats up, or will work nicely until it reaches a certain temperature and then fail. Even when an intermittent isn't specifically related to temperature, its frequently true that changes in temperature will aggravate the situation, thereby allowing you to find it more easily.

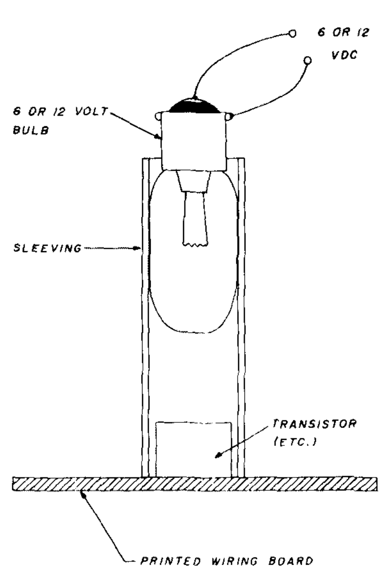

fig. 1. A simple device for heating individual components.

when the problem is heat

First, let's talk about how to heat up a set. Use a high-wattage lamp, sun lamp or hair dryer for general area heating to determine that the fault *is* temperature sensitive, rather than mechanical in origin. In a piece of equipment containing general power devices (or vacuum tubes), we can often heat up the circuits just by placing a box, towel, or blanket over the unit. This method is particularly useful for thermal faults that occur only in the cabinet. The thermal fault will continue for a few minutes after the box, towel, or blanket is removed, allowing time for troubleshooting.

Although area heating will give you the time needed to troubleshoot a fault, it won't help you find a specific thermally sensitive component. For this chore we must use local area heating. Several methods are available. A small high-intensity lamp, for example, will allow heating of a small area on a PCB. A soldering iron or gun will concentrate heat on an even smaller area almost to the exact component. (Be careful, — the hot tip of the soldering *tool can damage some* components, especially polyethylene capacitors.)

Another method used for heating individual components is shown in **fig. 1.** In this approach, the heat source is a 6 or 12-volt incandescent lamp. A No. 47 or No. 1891, for example. A small cylinder made of some material such as insulated sleeving ("spa-

ghetti") is designed to fit over components such as transistors and some integrated circuits. The heat source is placed in the open end, thereby concentrating the heat only on the component under suspicion. The tale will be told in about 30 seconds.

The indication that the component being tested is bad will be obvious. There will be a sudden change of operation, or a sudden increase in the noise produced by the circuit – the change is only rarely subtle.

fig. 2. Freeze spray will cool off components.

when the problem is cold

"Cold" can mean anything from room temperature to arctic temperatures. When equipment fails to operate in this range, area cooling is in order.

Area cooling is more difficult, in some cases, than area heating. Try putting small devices (up to, say, the size of a mobile transceiver) in the refrigerator for about an hour. I still fondly recall the looks on the faces of shocked boat radio customers when I placed their "won't work on cold days" VHF-FM transceivers in the shop refrigerator. (Many of those rigs are merely overpriced variants of 2-meter FM ham rigs, by the way). In most cases, 30 to 60 minutes in the refrigerator yields 5 to 10 minutes of troubleshooting time.

Local cooling is necessary for isolating components. Use a can of freon "freeze spray" as shown in **fig. 2.** (Electronic supply stores sell this product under several different brand names. The stores most likely to carry freeze spray are those whose clientele includes radio/TV/audio repair shops.)

Be careful not to spray too wide an area. Freeze spray is expensive and general area cooling won't help you find the bad components anyway. Use the spray only on individual components or small groups of components.

You can verify identification of the bad component by reheating it with soldering iron or the gizmo shown in **fig. 1.**

If the problem repeatedly appears and disappears on heating/cooling cycles, then you've found the source of the problem. Even if the problem isn't consistently repeatable, however, we can "work the odds" and replace the component "on speculation."

Our final method for fixing intermittents is another shotgun approach. In this case, however, we replace components on a "scattergun" basis. (*I can hear the howls from here! I admit it's not very elegant, and provides no balm at all to salve the ego of the technical genius. After all, any dumb grunt can unsolder a half dozen components and replace them. . . .* But let's consider some *facts of life.*)

I once worked in a hospital electronics laboratory that repaired clinical equipment. The emphasis was on low-cost, rapid repairs. One famous brand-name patient monitor used vintage circuitry. The ECG preamplifier and the DC power supply regulator used literally dozens of 2N3393, 2N3906, and 2N3904 plastic small-signal transistors. These transistors were typically connected six to eight at a time in circuits with multiple feedback and signal paths, all direct coupled. (Troubleshooting in circuits like this is a *dog.*) At that time, those transistors cost us $25 per hundred in bulk-packed bags. It takes 15 minutes to replace eight small transistors that cost $2 *total.* A total of 30 minutes put the equipment back on line.

The situation is only a little different for you. The biggest difference is that *you* buy transistors in overpriced blister packs rather than lots of a hundred; that's the price paid for buying onesie-twosies. Nevertheless, when the troubleshooting problem seems intractable, shotgunning components is a practical alternative!

You can be almost as efficient by removing components one at a time and testing each one as you go. If you find the faulty components, then it's a reasonable bet that the job is done. Unfortunately, the nature of intermittents – and Murphy's law – means that this form of troubleshooting frequently fails.

have a question for Joe Carr?

Send your question to Joe Carr, *ham radio,* Greenville, New Hampshire 03048. While not every letter can be answered personally, he will try to answer as many as possible in this column. – Ed.

ham radio

VHF/UHF WORLD

Joe Reisert, W1JR

Old Wives' tales and trivia

It's hard to believe that two years have passed since my first column appeared in *ham radio*.

Before I accepted K2RR's invitation, I spent much time trying to determine what kind of information would be required and how it should be organized. I knew that the material would have to be interesting, informative, easily understood, technically correct, and presented in proper sequence.

The selection of general topics was easier. After all, I knew that antennas, receivers, transmitters, propagation, and test equipment were invariably favorite subjects of conversation whenever Amateurs got together. So I narrowed these general topics down to a list of specific technical subjects I thought most VHF/UHFers and HFers would find useful and then tried to arrange them — in building block fashion — in such a manner that all the basics would eventually fall into place. This may explain why I didn't jump into highly technical subjects — such as microwaves — right away.

Finally, I drew on my vast file of letters, both those answered and those I hadn't had time to answer. These clearly identified both the subjects of greatest interest and those subjects that Amateurs find most confusing. This immediately flagged specific items within the more general list of topics that needed special attention. Some letter writers, of course, asked for articles about microwaves. They had to wait, since microwaves wouldn't be easy to discuss without a base to draw on.

Finally, the list of actual column topics and their tentative scheduling was complete. You've seen the result; in future columns, I'll try to cover new subjects, expand on subjects already covered, and explore further up into the microwave spectrum.

In the meantime, if you're following this series and can spare a few minutes, I need a favor: please drop me a note telling me what you liked and disliked about this column so far. Be frank. If I did a botch job on any subject, left you hanging on some item or left out a major point, let me know so I can try to correct the situation in a future column. I'll draw heavily on the letters for future topics.

Of course, as I've stated before, I can't possibly cover everything, nor can I answer all letters received. Just like you, I have only 24 hours in a day and a family that needs at least some of my time. After all, this is supposed to be a hobby!

So let's take a break from the usual format and see how sharp you are. *This will be your final exam on the first 23 installments and associated references listed therein.* With the help of some of my VHF/UHF friends and lots of letters to draw from, I've put together some trivia, some facts, some fallacies, and a few old wives tales. I hope they'll be fun to discuss and at the same time provide informative answers.

antennas

Let's start with everyone's favorite subject — antennas — and see what kind of old wives tales, etc., we hear. *Keep score, and no open books!*.

1. "If your antenna stayed up last winter it wasn't big enough." This is an original quote from one of the greatest VHF/UHFers ever, Sam Harris, ex W1FZJ, W1BU, KP4DJN, etc.

This statement is obviously not true. Maybe you live in Southern California. Or maybe you had a mild winter. Even better yet, maybe you built a big antenna but engineered the mechanics properly. It's nearly impossible to anticipate everything. How can any antenna survive a hurricane with winds exceeding 100 mph (161 kmph)? And how can you prevent your neighbor's tree from falling on and snapping your guy wires?

Tom, K8MMM, put it quite humorously in a recent letter: "If an outlandishly super-huge spectacle of an antenna doesn't stay up through *anything* nature has to afford for a particular area, it was too big, amateurishly conceived, and when down, due to all of the above — its owner is one of the least-heard-from and weakest things on the face of the earth!"

All things being equal, you can design a good antenna system with adequate gain if you follow the rules given in references 1, 2, and 3. Reference 4 discusses mechanical considerations and tubing strengths. *Do* build your antennas large enough, but not too large, and do so only with adequate mechanical strength!

2. "Always put your antenna as high as possible to get the best DX." This statement is basically true. However, there is a law of diminishing returns. First, if you're fortunate enough to be situated on a hilltop and have no obstructions, there's little to be gained by going up over 3 to 5 wavelengths. There's a problem if your

takeoff angle gets too low, especially when using F2, aurora, and sporadic-E propagation. At five wavelengths' height the takeoff angle will be about 3 degrees. Both EI2W and VE1ASJ found that often the optimum F2 signals come in with the antenna tilted upward several degrees! Ten wavelengths high is a real waste of time. If you do put up a large antenna on a high tower, it just may be large enough to meet the requirements spelled out by K8MMM.

Furthermore, especially on the UHF and microwave bands, the feedline loss can be horrendous.[5] Depending on frequency, going up an additional 50 feet (15 meters) may increase the effective gain by 1 dB while incurring another 2 dB feedline loss, an overall net loss of 2dB in station performance.[6] For VHF/UHF'ers I offer the following rule: *go high enough to clear local obstructions and STAY THERE. This is probably high enough*!

3. "Collinears are preferred over Yagis because they have broader bandwidth and larger capture area." This is truly a fallacy. While the bandwidth of a collinear antenna may be greater than a Yagi's, the matching method is often the limiting factor.[1] Also, what good is bandwidth when most VHF/UHFers never stray more than 50-100 kHz from the calling frequencies?[6]

As for capture area, this is a frequently misunderstood concept that was discussed in depth in reference 3. Capture area is directly related to gain regardless of the physical configuration of the antenna. Note that the capture area of a Yagi is usually much greater than its width and height, while with a parabolic dish it's the reverse, typically only 55 to 60 percent of the area of the dish.[3]

As discussed in reference 1, the choice of a collinear is primarily one of cost versus physical area. The collinear is usually a low-cost antenna, but takes up lots of area. The Yagi, while more critical to design, has less physical area and allows multiple antennas even on different bands on the same mast. The bandwidth doesn't usually enter into the final decision at all.

4. "Always stack Yagis two-thirds of the boomlength apart." Under certain conditions, typically with intermediate (2 to 3 wavelengths) Yagis, this may be true. However, for the vast majority of designs in use, and particularly the long Yagis, this would be a gross error.

This two-thirds boomlength rule was a common misconception when the effects of capture area were poorly understood. Stacking antennas too closely results in low gain, while stacking too far apart (especially with Yagis), gives high sidelobes and a beamwidth so narrow as to make aiming the antenna[3] properly very difficult. Check the beamwidth.[2] Then consult the charts in reference 3 and stack those antennas accordingly.

5. "The best Yagi designs are the ones produced by NBS." The NBS Yagis were a great stride forward. For once we had a cookbook with measured results. However, only six of the original NBS designs were provided in NBS Technical Note 688.[7] That represents only a small percentage of Yagi designs available to date.

The noteworthy item about the NBS Yagi antennas is that if they are properly duplicated, they will work to specification. However, if one of the six designs doesn't fill your requirements, or if you need a longer boomlength, there are no other NBS designs available.

Today there are literally an infinite number of other designs that will fit any length of boom desired.[8,9] Furthermore, using computer optimization techniques, up to 0.5 dB gain improvement is possible using existing Yagi designs such as the NBS 4.2 wavelength design.[10] We've just begun to open up a whole new area for improved Yagi antennas.

6. "T matches should be used on high performance Yagi antennas since Gamma matches don't work very well." This fallacy has been around for some time. It was fueled when all the new high-performance Yagi antennas, using balanced feeds and T matches, started springing up in the late 1970s. The Gamma match is capable of good performance. But it tends to inject a small imbalance into the design, which can cause a slight pattern skewing. The latter effect can be obviated by unbalancing the length of the driven element.[11]

Establishing a good ground for the Gamma return path is difficult. Hence, at UHF frequencies, the transmission line often gets hot and the VSWR and radiation pattern of the antenna changes as the feedline is moved to different positions. Furthermore, the size of the gamma rod can get out of proportion at the higher frequencies.

Therefore, a good T match with a built-in balun is hard to beat, especially when antennas are to be stacked. It takes more hardware, however, than a Gamma match and can be lossy if the balun design is not properly handled.

7. "Every time the size of an antenna array is doubled, the gain increases by 3 dB." This is true only in theory. Most antennas don't have a perfectly rectangular capture area or a clean pattern free of sidelobes.[3] As a result, antennas usually have to be stacked more closely than desired, with a resulting loss of gain.[3]

Stacking harness loss — which can approach 0.5 dB! — cannot be ignored. This is why the backplane feed system is recommended at the higher frequencies and where long transmission lines are used.[4]

Failure to provide sufficient mechanical strength not only in the individual antenna to be stacked but in the stacking frame can also cause gain to be lower than expected. Therefore, an array of long Yagis with a few feedlines is recommended over an array of smaller Yagis with more feedlines.[1] Plan on a 2.5 dB increase for each doubling of the array size.

8. "Sidelobes rob power and lower antenna gain." This is not necessarily true. The relationship of gain to sidelobe ratios was discussed in reference

2. It was shown in reference 3 that if the sidelobes are down 15 to 18 dB in the antenna to be stacked, the grating lobes in the final array should be down 13 dB for optimum gain. However, if the antennas to be stacked have sidelobes 13 or less dB down before stacking, they can't be optimally stacked. Each case must be studied separately, using references 2 and 3.

9. **"A good antenna requires a balun."** Not true. As stated above, a Gamma match can be effective. A well designed balun can do a great job of eliminating any radiation from a transmission line.[12] But an improperly designed balun, or one that uses a lossy transmission line, can actually *lower* the gain of the antenna! Proceed with caution and keep all balun losses as low as possible.

10. **"Front-to-side and front-to-back ratios are important antenna design parameters."** This statement is only partially true. The lobes at 90 degrees off boresight on a properly designed and built Yagi antenna are virtually at infinity in the E plane.[2]

A good front-to-back ratio may seem desirable for eliminating an interfering signal off the back of an antenna, but ratios exceeding 20 dB are not going to measurably improve gain or noise temperature. Furthermore, the angle subtended by the rear lobe is typically narrow so it is of dubious value. In a contest, it may be to your advantage to have some rear lobe radiation so you don't miss a new station or multiplier off the back of your antenna.

11. **"The more elements in a Yagi antenna, the higher the gain."** This is definitely false. Note that the NBS 4.2 wavelength Yagi has 15 elements, or two fewer elements than the 3.2 wavelength design. Yet it has more gain.[7]

What's more important in a Yagi antenna design is where the elements are placed with respect to each other (proper inter-element spacing) and the respective lengths of each director. There is a minimum number of elements to optimize the gain for each boom length. However, while extra elements may not be required, they can frequently be used to improve pattern and bandwidth of the design.[8,9]

12. **"Stacking improves gain and performance."** Properly executed, the gain of an array is improved if it's properly stacked.[3] Unfortunately, the operational performance may be degraded. For instance, if the beamwidth is too narrow in the horizontal plane, the array may not be optimum for meteor scatter operation, where the signals frequently arrive off the path.[13] Also, auroral propagation may be degraded by vertical stacking. The proper stacking, be it vertical or horizontal, is a function of the type of propagation desired.[6]

operating

1. **"The best place to operate is right on the calling frequencies since that's where the action is."** Unfortunately, there's plenty of truth to this statement, especially when good propagation conditions are occurring, such as during meteor shower and sporadic-E openings. This has been a real problem ever since the concept of VHF/UHF "calling frequencies" was instituted in the U.S.A. in 1978.[6] It's really sad, since it usually deprives all but the largest stations from sharing in the DX.

If the calling frequency concept is properly used, stations will call CQ or an appropriate station and *immediately* QSY to a different frequency so that others can then use the calling frequency. *Always remember to QSY at least 10 kHz away from the calling frequency.* QSYing only 5 kHz away very often causes QRM to other stations listening on the calling frequency, especially if adjacent stations are strong or if there is any splatter.

2. **"146.52 MHz is a good frequency for VHF contest operation since there are so many stations that operate there."** This freqency has been controversial for some time. In the early days of FM operation, most rigs were crystal-controlled and 146.52 MHz was often the only simplex crystal provided when the rigs were purchased. Hence it became a common meeting frequency for those who use simplex channelized FM. But as time went by, this frequency became very congested. What's more, it became a calling frequency for FM'ers as well as a frequency for passing emergency traffic. So it became a *real* sore point when contesting invaded 146.52 FM.

For the present, it's rather a moot question since contests — at least those run by ARRL — are no longer permitted on this frequency. Suffice it to say that this frequency should be used only as a calling frequency and for passing emergency traffic. Once contact is established, a quick QSY to one of the adjacent FM channels is suggested.

3. **"Scheduling stations during a contest, especially those stations out of your area, helps to improve your score."** This is probably not true unless you're a big contest station and desperately need every possible multiplier. Schedules do attempt to bring together people who normally may have difficulty casually running into each other. But schedules frequently take up valuable contest time. If too much time is used on schedules, there's a possibility that you may miss contacts by not working random stations that stay around for only a short time. Scoring must be carefully evaluated to see where your strengths and weaknesses lie.

4. **"Everyone should develop his or her own style of operating procedures."** This is an individual preference. It's always interesting when someone develops a new procedure that increases or improves communications. Such cases that quickly come to mind are meteor scatter and EME QSO procedures. But Amateurs who try to invent new procedures should be prepared to have lots of failures unless others know what they're up to! Confusion may follow when new routines or procedures are adapted. Signals on VHF/UHF are often weak, and any changes in operating procedures from

those normally used could cause confusion and incomplete QSOs.

propagation

1. "Operation on EME requires an investment of thousands of dollars." This is no longer true, especially if you're resourceful and are willing to build much, or even all, of your own gear. The most costly items associated with EME operaton are usually the antenna system and the power amplifier. There are now more than enough antenna designs available to enable you to "roll your own" EME antenna for any band where EME operation is presently conducted. Their performance can equal or exceed that of any commercially available antenna.

Likewise, there are plenty of designs available for power amplifiers.[14,15] There's no need to run expensive tubes unless you want to go to the legal limit. Many active EME'ers (including yours truly) have never run over 750 watts of output power and have been quite successful. You can also purchase or trade amplifiers.

Full legal power can definitely increase success ratios. Power helps when conditions are poor or if one of the EME stations is only marginal. (For further information, see references 16 and 17. The bibliographies at the end of these references provide more than enough information to help you keep costs down.) *The best advice I can give on keeping costs low is to try tested and proven designs. Avoid inventing new designs that may be more costly — especially if they're not successful!*

2. "The Perseids meteor shower always peaks on the morning of August 11." This is a myth. There are times every few years when this meteor shower *does* peak on this date and time. But you have to remember that the showers occur at the same time each year unless they're deflected by a planetary encounter. Since our year is 365.25 days long (that's why we have a leap year every four years), the shower, in Earth time, will occur approximately 6 hours later each year!

Even though the shower may peak at a specific time, that may not be the best time for a schedule, since the radiant* of the shower may not be in the proper location for communication in the direction desired. For example, it won't be very productive to operate when the radiant of the shower is on the other side of the earth, even if it *is* during the peak of the shower! (See references 13 and 18 for further information on this and other questions about meteor scatter communications.) At the end of each month's VHF/UHF World column I list the latest updated information on the predicted peak of the major showers; please note that this information does not include the location of the radiant.)

3. "Circular polarization is the only way you can operate on OSCAR." This has been proven false many times by those who regularly operate the satellites. It's true that circular polarization can yield up to 3 dB improvement on transmitting and receiving the satellites.

However, due to the geometry involved, the "sense" of circular polarization may actually reverse. As a result, you'd be significantly weaker on circular polarization during these times than if you used linear polarization if the sense reverses.

The bottom line is that you can use linear polarization on OSCAR with the possibility of a greater fading rate. Circular polarized antenna systems are recommended, *but only if they provide the capability for switching sense from clockwise to counterclockwise as required.*

4. "Good openings always occur with a high barometer." This is particularly true for tropospheric propagation, but is not true for aurora, F2, sporadic E, etc. In North America, the best tropo openings seem to occur during the spring near the Gulf of Mexico, during the summer on the California-to-Hawaii path, and during the summer and fall in the more northern latitudes.[6,19]

*Radiant — the point in the sky from which meteors appear to emanate.

Furthermore, the best tropo openings usually occur when a slow-moving high pressure (30.3 inches or 1025 millibars) area is present and mixed with warm moisture from the south. (For further information, see references 6 and 19.)

5. "The VHF/UHF bands are always open. It's just a case of no activity." This is a definite fallacy. I frequently hear this statement right after a VHF contest when the propagation conditions were good and there were lots of mountain-top stations. The dates of the ARRL June and September VHF QSO parties were purposely chosen to coincide with periods that have proven, over the years, to offer a high probability of extended openings. The August UHF and the spring Sprint contests are usually popular even though extended openings are few because of poorer propagation conditions at that time of year.

High locations do give some DX extension by virtue of the fact that they see a more distant horizon than a low-altitude station. Most mountain-toppers have shorter feedline and fewer obstructions to limit propagation. Although there are admittedly more mountain-top operations during the contests, there are fortunate persons who own mountain-top QTH's and are on the air year 'round. They can testify to the inaccuracy of the falsehood above.

Many good openings go undetected or are caught by only a few avid or lucky operators. Openings are missed mainly because of low, uncoordinated activity. We all know the frustration of calling a CQ with a highly directive antenna, only to find out later that a DX station was heard in our area, but that he had his antenna pointed in a different direction at the time of our CQ.

The best way to catch openings is to watch weather maps, be vigilant during the most likely seasons for openings[6,19] monitor the calling frequencies, take advantage of propagation beacons, and adhere to nightly schedules as well as uniform activity nights and hours.[6]

6. "It takes a hurricane to get a good tropo opening." I have long observed that the good openings at higher latitudes often occur when hurricanes occur south of the path (references 6 and 9). But there are exceptions — for example, the Gulf of Mexico opening in the spring, and openings in the fall in the more northern latitudes.

What's required for an opening is explained in item 4 above. Yet the coincidence of longer DX openings occurring when hurricanes are present cannot be denied. Hurricanes cause low pressure areas to develop. These low-pressure areas affect high pressure areas, causing them to build up and move slowly — this slow movement, combined with the warm moisture drawn from the low pressure area, results in extended openings.[6]

receivers

1. "You need a GaAs FET preamp to work DX on the VHF/UHF bands." This is definitely false. How do you explain all the DX before solid-state devices were available in the 1960s? There are plenty of good JFETs, MOSFETs, and bipolars that yield low noise figures (1 to 2 dB), which is more than sufficient for non-EME modes where local noises are frequently the limiting factor in communications.

There's no denying that GaAs FETs are becoming very popular.[20] In many cases, they've improved receiver sensitivity. But the fallacy that they're the *only* devices that work well has probably been irrevocably spread since antenna-mounted preamplifiers are now quite popular and most use GaAs FETs.

2. "A low noise figure receiver will always outperform one with a high noise figure." You can't deny that receiver sensitivity is a great factor in communications. After all, "If you can't hear them, you can't work them." But noise figure is only one part of the equation. High dynamic range is also extremely important, especially if other strong stations are present. Low-noise preamplifiers often have poor dynamic range.[20] High gain ahead of a mixer, especially a poor one, can cause blocking and IMD as well as other annoying phenomena.[21]

Many modern receivers, and in particular the synthesized HF transceivers, have very poor phase noise and are easily overloaded.[22] Therefore, in order to hear the weak ones, attention must be paid to the dynamic range as well as the noise figure of the receiver.

3. "28 MHz is a good IF for a VHF/UHF converter." Generally speaking, this is true. Attention must be paid to dynamic range, as just discussed. However, when you go up into the UHF frequencies, image rejection becomes a real problem.[23] It's not a trivial problem to filter out an image only 56 MHz away from a 1296 MHz converter without incurring some undesirable filter loss.[24] Image rejection or image recovery mixers are recommended.[23] For simplicity, 2-meter (144-144.5 MHz) IFs are an acceptable alternative and are becoming very popular.

4. "A 1 MHz crystal calibrator makes an accurate VHF/UHF frequency calibrator." This is definitely false. Crystals below 3 MHz (because of their "cut") are usually much less stable than those between 3 and 10 MHz.[25] Furthermore, 3 and 4 MHz markers are very convenient because they can place loud, easy-to-find calibration points in most receivers on 144 MHz and above.

The calibrator discussed in reference 25 is highly recommended and should be an essential part of every well-equipped VHF/UHF station regardless of the equipment used. Your success rate drops dramatically when you don't know your frequency within at least 1 kHz, especially on EME and meteor scatter communications.[13,16,17]

5. "You need a Hewlett Packard 8970A noise figure meter to accurately tweak a preamplifier to its lowest possible noise figure." This is definitely false. There were plenty of optimized low-noise preamplifiers long before the HP 8970A arrived on the scene a few years ago. The older AIL models 74 and 75 as well as the HP 340 series gave good results.

The principal reason for the popularity of the HP 8970A is that it's quick and easy to use, measures both noise figure and gain simultaneously, and has a digital readout with 0.01 dB precision. It's been shown, however, even by the manufacturer, that the results are probably only accurate to ±0.5 dB. It's also been shown more recently that preamplifiers with poor input VSWR (such as most GaAs FETs) can cause large measurement errors (up to 0.5 dB) if you don't use a noise tube with at least 10 dB of extra internal attenuation (such as the newer HP 346A type).[20]

Don't be fooled by digital readouts and extravagant claims. What's really important is *whether your preamplifier is optimized to the minimum noise figure it can deliver* (this can be done with either the older or newer instruments) and *how it stacks up with other designs* (comparison at noise figure measurement parties).

6. "A 1N21 or back biased transistor makes a good noise figure generator." This is usually untrue. Many of the older 1N21 type of noise generators had terrible VSWR that caused the preamplifier under alignment to be optimized to the noise generator impedance rather than to 50 ohms. This can also be true with transistors. Always use at least 10 dB of attenuation, preferably an attenuator "pad" with low (1.2:1 maximum) VSWR, between your noise generator and the device-under-test. (See item 5 above.)

7. "A Dow-Key relay won't have sufficient isolation at VHF/UHF frequencies." This can be misleading unless the specific type of relay is stated. Some of the Dow-Key relays have an extra isolation feature. Other manufacturers also have isolation problems.

It has been pointed out that for safety's sake, the power entering a low-noise preamplifier should not exceed 10 milliwatts and 100 milliwatts at worst case.[16] It was also shown in

reference 16 that for optimum results and safety, a two-relay system is recommended. The second relay should return the preamplifier to 50 ohms during transmit to prevent amplifier oscillation and possible destruction during transmission periods.

It was further pointed out in references 16 and 26 that the length of transmission line between the two relays is important if the increased isolation is to be obtained. Suffice it to say that much more attention should be paid to the relay types used and at least 50 dB of transmit-to-receive isolation is highly recommended, especially when high power is used.

8. "I built it just like the article and it didn't work." Oh, how often authors hear this statement! For this very reason I've often spent hours carefully writing and then *rewriting* my column to make sure that everything is perfectly clear. Proof copies are carefully scrutinized several times through the various stages of production. However, bugs do occasionally creep in!

Suffice it to say that all circuits should be duplicated *exactly* as shown (providing that an error hasn't crept into the schematic!) unless you have enough test equipment and experience to outwit the author. I must admit that I will sometimes not publish new designs for fear that they may be too complicated or will be likely to cause a rash of angry letters. *If you alter an author's circuit, however slightly, or substitute a different part than specified, don't blame the author or ask him for help if you experience a problem!*

transmission lines

1. "Open-wire transmission line has less loss than coax." This is probably true if the VSWR on the open wire line is low, or if there's no contamination or moisture on the insulators.[4] However, open wire lines must be relatively straight and be kept away from other lines and antennas. As a result, coaxial cable, even though it may have slightly higher loss, may be more desirable, especially when multiple antennas and feedlines are present on the same mast. (For further information on this subject, see references 4 and 5.)

2. "Always cut phasing lines in multiples of one-half wavelength." This theory was debunked in reference 4, where it was pointed out that for proper power distribution, odd numbers of *quarter-wavelength* feedlines are preferred. (Refer to reference 4 and its references for further information on this subject.)

3. "The way to improve EME antennas is to replace all coax with open wire lines." This subject was discussed in detail in references 4 and 5. For many of the reasons mentioned above, coaxial cable, properly chosen and used, may be preferable to open wire line.

4. "You need a Bird wattmeter to accurately measure transmitter output power and VSWR." This is also not true. There are other suppliers of good accurate power/VSWR meters. Most power meters have their own limitations. For instance, the *accuracy* of the power indicated is usually only ± 5 percent of full scale. This means that a 5-watt error is possible on the 100-watt scale. This can really affect the power measurement at a 25-watt power level on the same scale!

VSWR measurement accuracy is affected by the directivity or ability of the instrument to be able to distinguish between a true and a poor VSWR. Typically 20 to 30 dB is the limit, meaning that VSWR readings below 1.2:1 may be inaccurate.

Accurate readings of VSWR can be accomplished at low levels using the techniques and inexpensive coupler or VSWR bridge described in reference 27. If you build a hybrid coupler similar to the one in reference 27, you can build your own power meter and calibrate it against a borrowed meter. Unless you're measuring power near the legal limit or are trying to measure the *efficiency* of a high-power amplifier, an expensive power/VSWR meter is not required. But once you use one, you'll be hard pressed to do without it.

5. "A 1.5:1 VSWR is good enough." This is true. But where is the VSWR measured, and how accurate is the VSWR meter? Reference 5 pointed out that the length and loss of the transmission line between the antenna under test is extremely important on VHF and higher frequencies. For instance, a line loss of 7 dB (not uncommon on some of the higher bands where long runs are needed) transforms an open or short circuit (infinite VSWR!) at the far end into 1.5:1 VSWR at the near end.[27]

It can't be stressed enough that for optimum performance on the VHF and the higher bands, the quality of the VSWR meter as well as the feedline loss[5] must be taken into account when testing for VSWR!

6. "RG-58 can be used on 432 MHz." True — but the results may be disastrous! This type of line normally has a loss of over 10 dB per 100 feet (30.5 meters) and can handle only about 75 watts safely at 432 MHz.[5] So RG-58 coax cable should be used only sparingly in places where the line loss is not critical.

7. "PL259s are OK at 432 MHz." True. But this is so only if the PL259 is properly integrated with the coaxial cable. It must be stressed that the PL259 is not watertight, doesn't have a guaranteed VSWR, and probably can't handle much power on the UHF frequencies. Therefore, it should be avoided if at all possible.

8. "Heliax™ and hardline transmission lines are too expensive for Amateurs." This is a common misconception.[5] The cost of generating high power at VHF/UHF frequencies and high transmission line losses are usually the limiting factors in successful communications. Placing preamplifiers at the top of a tower helps the receive path, but transmitters (especially the high power tube type) are not readily mounted at the top of a mast.

Good quality feedline such as Heliax and hardline, with their low insertion

loss, will not only deliver the most "bang for the buck," but will also frequently outlast lower cost transmission lines by a 2 to 5:1 ratio. If remote relays are used, they can do double or even better duty by servicing multiple antennas. As a result, *the high initial cost is quickly amortized over the years and the performance is top notch to boot.* Couple this with the favorable prices often found at flea markets and you have a super bargain!

9. "The G-line is a nickname for the chorus line at a burlesque house." I ask this question to see if you're still awake. "G-line" was mentioned in my October, 1985, VHF/UHF World Column.[6] It's basically a single wire transmission line similar to a toy "string telephone." It has many exciting possibilities for low loss and inexpensive installations. See the October article for further information.

transmitters

1. "VHF/UHF amplifiers typically have 20 dB of gain." This theory, which has been around for some time now, invariably causes grief when someone discovers that you can't run 1500 watts of output with a 10-watt driver.[14] Typically speaking, the gain of most VHF amplifiers is 15 to 20 dB and 10 to 16 dB at UHF. This is only if the amplifier is operated in linear service. Class C has lower gain and is not recommended for reasons mentioned in reference 14.

The more modern grounded grid triodes frequently have 3 dB lower gain than this although they are usually more stable than the older neutralized designs. When you buy or build an amplifier, check the specifications beforehand and see what the drive requirements will be. You may need an additional driver to get the output power expected from your amplifier.[28]

2. "You can run a single 4CX250B at 500 to 600 watts output." True, *but your tube won't last very long*! Amateurs seem to have a thing about running devices past manufacturers' ratings! They frequently provide insufficient cooling to boot. Better read references 14 and 15 and drop your power, too. Both you and the tube will be friends for a longer time!

3. "You can't operate 2-meter EME without an 8877 amplifier." If there ever was a misconception, this is it. *Hundreds* of Amateurs operate EME without running the legal limit or using a high power tube like the 8877.

Other tubes that will deliver the same power[14,15] are available. Tubes can also be run in parallel. You can operate EME with as low as 500 watts of output power if you're patient, have sufficient antenna gain, and "have your act together."[16,17]

4. "Speech processing extends your SSB transmitting range and prevents you from splattering." This statement is true as long as you use the speech processing properly and don't overdrive your transmitter. All too often, Amateurs not only run their amplifiers in a highly non-linear fashion but also run improperly adjusted speech processors. If you run speech processing, the duty cycle on the power amplifier will increase. If you don't increase cooling to the final stage, you could experience premature failure.

5. "If your maximum output power level is 100 watts you should occasionally see that power level register on SSB peaks on your output power meter." Boy, here's another big lie. Ever since reasonably priced power meters became available, they've been used to do the wrong things. For starters, most power meters have a highly damped meter movement. As a result, they respond slowly. The truth of the matter is that under normal circumstances, the power meter should be indicating no more than about 25 to 30 percent of the actual power level of the amplifier in a key-down position.[28] It's for this very reason that solid-state amplifier/drivers have gotten such a bad reputation.

6. "You can run a pair of 4CX250B's at 1 kW output on SSB." This is true. However, you won't have many friends. The 4CX250B is rated on SSB operation at 500 watts input for 300 watts PEP output per tube with an IMD of only 25 dB.[14] IMD of 30 dB is considered Amateur standard. So don't argue with fellow Amateurs when they say you're splattering and you are only running 600 watts PEP output from a pair of 4CX250Bs!

7. "IMD isn't a real problem on the VHF/UHF bands since there are so few stations and they're all geographically separated." Amateurs on VHF/UHF used to say that they didn't have to worry about dirty signals on VHF/UHF since there were so few people and so much spectrum available. But that's all changing now, with many stations coming on and often operating in close proximity to the calling frequencies. Line of sight, higher gain antennas, and sensitive receivers that often lack high dynamic range are compounding the problem.[21,22] Better start watching your signal quality as closely as you do on the DC bands.

summary

This month's column, sort of a mixed bag, was intended to put to rest several of the most popular old wives' tales. Most of the statements made were addressed in the past 23 columns and the other references cited. I hope you'll be sufficiently interested in the subjects discussed to research the referenced material independently. At the same time, I hope you've enjoyed this departure from my usual format and will continue to follow this column as faithfully as you have. Don't forget to drop me a line with any suggestions or advice.

acknowledgements

I'd like to thank all the unnamed Amateurs who brought the statements for this month's column to mind. Unfortunately, I can't name you all, but some of you would rather not be identified anyhow! I do want to extend special thanks to Lewis Collins, W1GXT, and Gary Madison, WA2NKL, for helping me assemble many of the statements used as the basis for this month's column.

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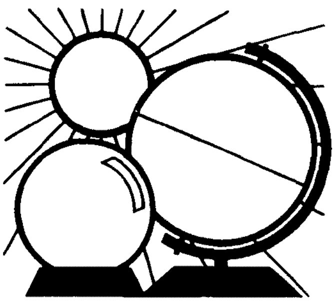
important VHF/UHF events

December 2:	*7 to 11 PM local SWOT 2-meter QSO party (information from K5IS)*
December 11:	*EME Perigee*
December 13:	*Peak of the Geminids Meteor Shower predicted at 0650 UTC*
December 21:	*Peak of the Ursids Meteor Shower predicted at 2200 UTC*
December 21:	*±1 month, winter peak of sporadic-E propagation*
January 3:	*Peak of the Quadrantids Meteor Shower predicted at 2300 UTC*
January 8:	*EME perigee*
January 11-12:	*ARRL VHF Sweepstakes contest*

ham radio



DX FORECASTER

Garth Stonehocker, KØRYW

MUF Forecasting

Recent studies of variations in maximum usable frequency (MUF) and its controlling factor, foF2 (the maximum ion density of the ionosphere), show how to do MUF forecasting (a day or two ahead) during the next year of DXing. In previous columns, methods for obtaining a mid-latitude noontime foF2 or MUF baseline (average) value using the average solar flux value for the same month were provided.[1]

The daily percentage change in MUF can be obtained by using a factor related to the daily change in solar flux or geomagnetic A index. The factor given in 1984 was percentage change in MUF equals 30 per cent of the solar flux change (1 per cent for every 3 flux units). This was when the solar flux numbers were in the 150s, with large daily excursions of 10 to 20 units and a 2 to 3-day delay for the ionosphere's foF2 to catch up. This approach to forecasting really works!

Now that we're near minimum sunspot number, does the same factor still apply? No. Since the ionosphere is a geophysical system in equilibrium (i.e., balanced), expect compensating conditions to occur even though MUFs are lower. The large and fast variations no longer occur but are slower-changing — 10 to 20 units in value over a period of several days.

One study shows that the ionosphere is now more sensitive to solar flux changes. The new factor has each flux unit equal to 1.2 the percentage foF2 change. In addition, the 2 to 3-day delay is no longer experienced because solar flux variations occur slowly enough for the ionosphere to "keep in step."

The study also indicated that the influence of geomagnetic field variations on the ionosphere is greater now than when solar flux levels were higher. For example, at higher fluxes, an A of 16 to 30 decreased the foF2 by approximately 4 to 7 per cent and an A of over 100 resulted in a 15 per cent decrease from the median value of the month. At the current flux level, an A of 11 to 70 causes an 8 to 25 per cent decrease in foF2. There appears to be quite a difference. However, the foF2 median value at the higher value of flux was 9 MHz; currently it is 5.5 MHz. Take 15 per cent of 9 MHz and 25 per cent of 5.5 MHz. Notice that these values are very close to 1.25 MHz — the actual foF2 reduction — in either case.

The ionosphere has a way of equalizing effects between sunspot extremes. These foF2, solar flux, and A index relationships were further confirmed by a study conducted at the Institute of Telecommunication Sciences (ITS) in Boulder, Colorado. The study examined the distribution of MUF values about the median (value) for the month over a 6-hour period during each day during the various seasons and over three sunspot number ranges. They found little difference (only 2 per cent) between mid-latitude MUF variation and sunspot levels. (More on this study next month).

In summary, the current solar flux increases, though small, cause the mid-latitude MUF to increase (1 per cent for each flux unit). During disturbed geomagnetic field conditions, the reduction in mid-latitude MUF (which occurs a few hours after the onset of a storm) can be found from this relationship: percentage change in MUF = $0.375A + 3.75$. Values of solar flux and geomagnetic field indices are broadcast by WWV at 18 minutes after the hour. These new factors should help you more accurately forecast DX conditions during the next few years of low sunspot numbers.

last minute forecast

The higher HF bands (10 through 30 meters) are expected to be best during the first and second weeks of December as well as part of the last week of the month. A solar flux peak on December 5 and another on January 1 are expected to occur, enhancing DX conditions. Lower solar flux values will mean lower MUFs during the third and fourth weeks of December. However, lower flux means greater daytime signal strengths on the lower HF bands since there'll be less absorption during these times. Lower absorption normally occurs during the winter months as well as during the 27-day solar cycle minimums. The geomagnetic field will probably be disturbed during the third week of the month. These disturbances result in a reduced MUF on east-west and northern paths and an enhanced MUF on transequatorial paths.

The Geminids meteor shower, which will peak on December 13-14, will provide the richest and most reliable display of the year, with rates of 60 to 70 per hour. Because optical observations may be difficult or impossible during periods of poor weather in December, actual numbers must be determined by radio reception. A smaller version of the shower will be observed on December 22.

Lunar perigee and a full moon will occur on December 11 and 27, respectively. Winter solstice occurs on the 21st at 2208 UT.

DECEMBER

WESTERN USA

GMT	PST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
0000	4:00	40	40	20	12	10	12	10	20
0100	5:00	30	40	20	15	10	12	10	20
0200	6:00	30	40	20	20	12	12	12	20
0300	7:00	30	40	20	20	20	12	15	30
0400	8:00	30	40	20	20	20	12	20	30
0500	9:00	30	*40*	20	20	20	12	20	30
0600	10:00	40	*40*	20	20	20	15	20	40*
0700	11:00	40	*40*	*30*	20	20	20	20	40
0800	12:00	40	*40*	*30*	*30*	20	20	20	40
0900	1:00	40	*40*	*30*	*30*	20	20	*30*	40
1000	2:00	40	*40*	*30*	*30*	*30*	20	*30*	40
1100	3:00	40	*40*	*30*	*30*	*30*	*20*	*30*	*40*
1200	4:00	40	*40*	*30*	*30*	*30*	*20*	*30*	*40*
1300	5:00	*40*	*40*	20	20	*30*	*30*	*30*	*40*
1400	6:00	*40*	30	12	12	20	*30*	*30*	*40*
1500	7:00	*40*	20	12	10	15	*30*	20	*40*
1600	8:00	40	20	10	10	15	*30*	20	*40**
1700	9:00	40	30	10	10	12	20	20	*80**
1800	10:00	*80*	30	10	10	12	20	20	*80*
1900	11:00	*80*	40	12	10	12	15	20	*80*
2000	12:00	*80*	40*	15	10	12	15	15	30*
2100	1:00	*80*	40*	20*	10	12	12	15*	20
2200	2:00	*80*	40	20	10	12*	12	12	20
2300	3:00	*80*	40	20	10	10	12	12	20

MID USA

MST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN	CST
5:00	40	40	20	20	10	12	10	30	6:00
6:00	40	40	20	20	12	12	12	30	7:00
7:00	30	40	20	20	20	12	15	30	8:00
8:00	30	40	20	20	20	20	20	40	9:00
9:00	30	*40*	20	20	20	20	20	40	10:00
10:00	30	*40*	20	20	20	20	20	40	11:00
11:00	40	*40*	*30*	20	20	20	20	40	12:00
12:00	40	*40*	*30*	*20*	20	20	20	40	1:00
1:00	40	*40*	*30*	*30*	20	20	*30*	40	2:00
2:00	40	*40*	*30*	*30*	*30*	*30*	*30*	40	3:00
3:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	40	4:00
4:00	*40*	30	*30*	*30*	*30*	*30*	*30*	*40*	5:00
5:00	*40*	20	15	*30*	*30*	*30*	*30*	*40*	6:00
6:00	*40*	20	12	15	*30*	*30*	*30*	*40*	7:00
7:00	40	20	10	12	20	*30*	20	*40*	8:00
8:00	40	20	10	12	15	20	20	*40*	9:00
9:00	40	20	10	10	15	20	20	*40*	10:00
10:00	40	20	10	10	12	15	20	*80**	11:00
11:00	40	30	10	10	12	15	20	*80*	12:00
12:00	*80*	30	12	10	12	10	20	*80*	1:00
1:00	*80*	40	15	10	12	10	15	30	2:00
2:00	*80*	40	20	10	12*	15	12	20	3:00
3:00	*80*	40	20	12	12*	12	12	30	4:00
4:00	*80*	40	30	15	10	12	12	30	5:00

EASTERN USA

EST	N ASIA FAR EAST	NE EUROPE	E S. AFRICA	SE CARIBBEAN S. AMERICA	S ANTARCTICA	SW NEW ZEALAND	W OCEANIA AUSTRALIA	NW JAPAN
7:00	40	40	20	20	20	15	15	40*
8:00	40	40	20	20	20	20	20	40*
9:00	40	40	20	20	20	20	20	40
10:00	40	*40*	20	20	20	20	20	40
11:00	40	*40*	20	20	20	20	20	40
12:00	40	*40*	*30*	20	20	30*	20	40
1:00	40	*40*	*30*	20	20	30	20	40
2:00	40	*40*	*30*	*20*	20	30	*30*	40
3:00	*40*	*40*	*30*	*30*	*30*	30	*30*	40
4:00	*40*	*40*	*30*	*30*	*30*	*30*	*30*	40
5:00	*40*	*30*	15	*30*	*30*	*30*	*30*	40
6:00	*40*	20	12	15	*30*	*30*	*30*	*40*
7:00	30	20	10	15	*20*	*30*	*30*	*40*
8:00	40	20	10	12	20	20	20	*40*
9:00	40	20	10	12	12	20	20	*40*
10:00	40	20	10	12	12	20	20	*40*
11:00	40	20	10	10	12	20	20	*40*
12:00	40	20	10	10	12	15	20	*40*
1:00	40	30*	10	10	12	12	20	*80*
2:00	*80**	30	12	10	12	12	20*	*80*
3:00	*80**	40	15	10	12	10	15	*80*
4:00	*80*	40	20*	10	12*	10	12	20
5:00	*80*	40	30	12	12*	10	12	20
6:00	*80*	40	30	15	12	12	12	30

The italicized numbers signify the bands to try during the transition and early morning hours, while the standard type provides the MUF during "normal" hours.

*Look at next **higher** band for possible openings.

band-by-band summary

Ten and twelve meters, the highest day-only DX bands, are nearest the MUF for southern hemisphere paths. They will be open most days when the solar flux is above 75 during the 7 to 10-hour period centered on local noon. These bands open on paths toward the east and close toward the west. The paths are up to 4000 km (2400 miles) in single-hop length, and on occasion double that during evening transequatorial openings.

Fifteen meters, a day-only DX band open most of each day, has lower signal strengths and greater multipath variability than 10 and 12 meters. It will be best when the MUF is just resting above this band, until it drops below it — a transition period that occurs right after sunrise and just before sunset. Transequatorial openings will occur, with distances similar to 10 and 12 meters.

Twenty, thirty, and forty meters are both day and nighttime DX bands. Twenty is the maximum usable band for DX in the northern directions during daytime. In combination with 30 meters, it provides nighttime paths for the day-only bands. Forty meters becomes the main over-the-pole DX daytime band, with some hours covered by 30 meters. This path may be affected by anomalous absorption during a few days of the month.

Eighty and one sixty meters, the night-only DX bands, exhibit short-skip propagation during daylight hours, then lengthen at dusk. These bands follow the darkness path, opening to the east just before local sunset, swinging more to the north-south near midnight, and ending up in the Pacific areas for a few hours before dawn. On some nights, 80 meters, with its high signal strengths, will be the best band to use. One-sixty is also expected to provide good conditions. Please remember the DX windows of 3790 to 3800, 1825 to 1830, and 1850 to 1855 kHz.

references

1. Garth Stonehocker, KØRYW, "DX Forecaster," *ham radio*, January, 1984, page 83, and February, 1984, page 79.

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REVIEW

Alpha Delta's multi-band twin sloper antenna

Slopers have enjoyed considerable popularity over the past few years. Recently a number of different manufacturers have been producing several models of multi-band slopers that are fairly convenient to install and apply and give a pretty good accounting of themselves on the air.

The newest sloper is the DX-A from Alpha Delta, designed to cover 160, 80, and 40 meters. Alpha Delta is well known for its MACC power supply switch and transient protector and its Transi-trap antenna lightning protector. Don Tyrell, W8AD, president of Alpha Delta, was looking for a way to improve his low band signals and came upon the March, 1981, *QST* article by Doug DeMaw, W1FB, "More Thoughts on The Confounded Sloper." Don researched the sloper further and decided that the design could give him the kind of performance he was looking for.

Because he wanted a multi-band antenna, he had to make a few modifications. The first was to modify the basic design to configure it more like an inverted dipole rather than a single wire antenna (as described in the *QST* article and produced by other manufacturers.) This was done to broaden the bandwidth and improve the radiation efficiency of the antenna. As first designed, the antenna had one element that covered 160 and 80, while the other tuned 40 meters. However, after a number of these units were produced and out in the field, Don found that the antenna's performance could be improved if the 160-meter resonator was placed on the end of the 40-meter wire (see **fig. 1**).

Since I had the original DX-A antenna, I had to make a number of minor modifications before this review could begin.

Alpha Delta uses an aluminum tower bracket drilled to fit a Rohn tower bolt and has an SO-238, female UHF connector to simplify attaching the antenna feed line. The two antennas are fed from this common point and extend away from the tower just as a dipole would. If there's any question of a good ground connection, such as with a crank-up or older and possibly corroded tower, it will be necessary to run a grounding wire to ensure proper operation.

Alpha Delta recommends that the DX-A be placed between 25 and 40 feet up — 30 is suggested as optimum. The elements should be run as closely to 180 degrees apart as possible. (The test antenna was installed at 32 feet on a 56 foot

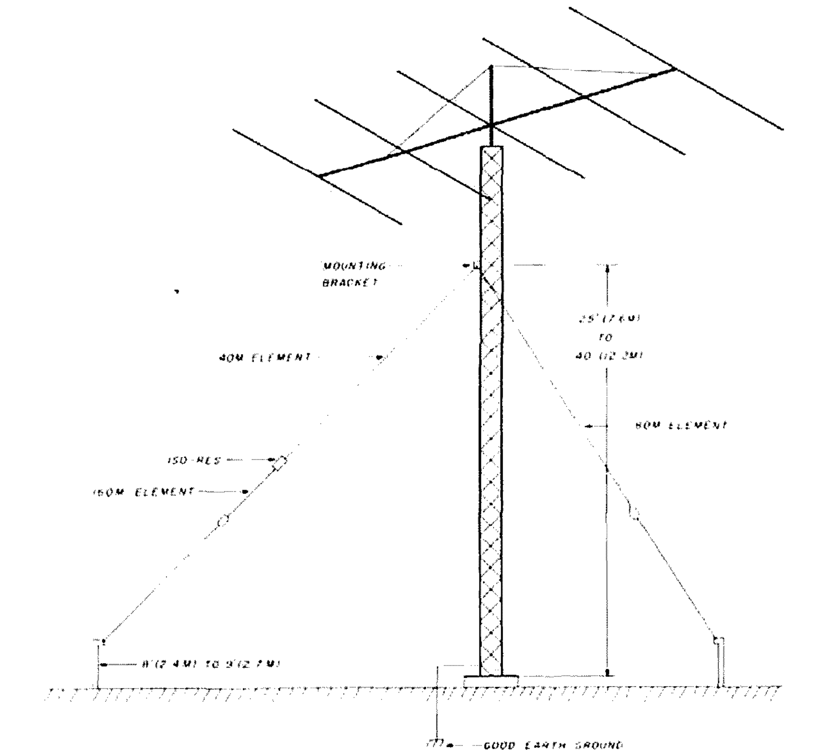

fig. 1. Performance was improved by placing 160-meter resonator at the end of 40-meter wire.

tower.) Other dimensions can be used without seriously degrading performance.

The antenna is a single-wire radiator for 80 meters and a loaded radiator on 40 and 160. It uses a low-Q "iso-resonator" instead of an L-C trap to eliminate the problems of capacitor break-down under high RF loads. This coil is rated at 1500 watts PEP. Don's new dimensions for the antenna are:

80 meters	67-69 feet
40 meters	34 feet
160 meters	18 feet

Tuning 80, I found that I was able to get an acceptable match at 3.6 MHz without too much difficulty. However, finding that the suggested dimensions for 40 meters were a bit too long, I spent a considerable amount of time pruning the antenna in an attempt to get the antenna to resonate. I was finally able to get the unit to cover the whole band when the antenna was approximately 32 feet long. Luckily at this point, the antenna was just about resonant at where I wanted it for 160. One cut and it was at 1.84 MHz.

Should you have any problems during installation or tune up, you can call Alpha Delta's technical department at 513-376-4180. Their representatives will be able to provide you with sound advice.

Less than 2:1 SWR bandwidth for this installation is:

160 meters	50 kHz	1.815-1.865 MHz
80 meters	200 kHz	3.5-3.7 MHz
40 meters	300 kHz	7.1-7.3 MHz

This antenna can also be tuned with open wire feeders or coax and a transmatch for greater frequency excursions. As modified, the DX-A will also tune the 30 and 17-meter Amateur bands with a reasonable SWR.

It is far too soon to tell how well this antenna will perform on long-haul DX contacts. However, early results indicate that it will be pretty good on 80 and 40 with credible results on 160.

The multi-band sloper won't give you the performance of full-sized antennas. But, if your space is limited or if you need a single antenna to get on the air, the DX-A is a perfect candidate for you to consider.

For more information, stop by your local dealer or contact Alpha Delta, P.O. Box 571, Centerville, Ohio 45459.

Circle #301 on Reader Service Card.

N1ACH

packet goes portable

GLB Electronics, Inc., has introduced the first packet radio controller designed for portable and solar powered stations.

(continued on page 124)

NEW products

Unique in terms of small size and low current drain, the PK1L is designed for portable use and for remote digipeater operation. At a current drain of only 25 mA, the PK1L can even be operated on a 9-volt transistor radio battery. PK1L also has self-contained "watchdog" and power-down sensing circuits with a lithium battery backup for memory. When power is disconnected all parameters and modes that have been set up are retained and become available again when power is restored. For remote digipeater operation in the event of malfunction due to nearby lightning strikes, etc. the watchdog automatically resets the CPU, ensuring continued operation for anything short of physical damage to the unit. The PK1L also provides a "connected" signal, plus two spare inputs and outputs that can be programmed for custom applications. A "remote command lockout" input can be used to prevent unauthorized stations from sending control commands.

The PK1L is housed in a rugged, all-metal, shielded enclosure measuring only 4.6 × 5.9 × 1.0 inches. The PK1L is entirely self-contained with an onboard CPU, 8K of memory, preprogrammed 32K ROM, RS-232 interface and packet MODEM weighing only 12 ounces. Both connectors are DR-25s, chosen for ready availability. Pinouts were chosen to preclude damage due to improper insertion. One connector is for transceiver and power supply and the other for a computer, a computer terminal, or a teletype machine, either ASCII or Baudot. The CPU is a CMOS Z80A microprocessor operating at 3.58 MHz.

For further information, contact GLB Electronics, Inc., 151 Commerce Parkway, Buffalo, New York 14224.

Circle #303 on Reader Service Card.

ICOM IC-R7000 receiver

A new continuous-coverage receiver from ICOM monitors all Amateur Radio frequencies, from 25 MHz through 2000 MHz in FM, AM, and SSB modes. (Specifications guarantee from 25 to 1300 MHz.) The unit also covers aircraft, marine, government, emergency services and television bands.

Nintey-nine memory channels are featured. Frequencies are accessed by either keyboard or tuning knob. Scanning speed is adjustable. Five tuning speeds (0.1 kHz, 1.0 kHz, 5 kHz, 10 kHz, 12.5 kHz, and 25 kHz) are available. The fluorescent display includes a dimmer switch for comfortable viewing.

The compact unit measures 4 3/8 x 11 1/4 x 10 7/8 inches and is priced at $899. Infrared remote controller and voice synthesizer are optional.

For details, contact ICOM America, Inc. 2380 116 Avenue N.E., Bellevue, Washington 98009-9029.

Circle #304 on Reader Service Card.

AZDEN 2-meter transceiver

Amateur-Wholesale Electronics has announced the new AZDEN PCS-5000 2-meter microcomputer FM transceiver.

The PCS-5000 has an unprecedented frequency range of 140.000-152.995 MHz, allowing the unit to be used for CAP and all MARS frequencies. Its small size — only 2 inches high by 5-1/2 inches wide by 7-1/4 inches deep — allows the radio to be placed almost anywhere.

The microcomputer facilitates features not previously available, including up to 11 nonstandard splits, 20 channels of memory in which offset and PL information can be stored, dual memory scan, scan lockout in memory mode, two ranges of programmable band scanning, with selectable scan increments, busy scan and delay scan in both the memory and band-scan modes, discriminator scan centering (AZDEN exclusive patent), priority memory with alert tone, state-of-the-art lithium battery for memory backup, repeater reverse, acquisition tone, programmable PL generator, and direct frequency entry.

The crisp, easy-to-view backlighted liquid crystal display shows operating functions as well as frequency and S/RF bar-graph meter. The keyboard is backlighted for easy viewing even in total darkness.

Other features of the PCS-5000 include high/low power (25 watts and 5 watts, fully adjustable), a superior receiver with *unprecedented* sensitivity and dynamic range, true frequency modulation, 16-key touchtone pad, a rugged multi-function dynamic microphone, a built-in speaker, mobile mounting bracket, remote speaker jack, and all cords, plugs and fuses.

For further information, contact Amateur-Wholesale Electronics, Inc., 8817 S.W. 129 Terrace, Miami, Florida 33176.

Circle #305 on Reader Service Card.

new Heathkit catalog

More than 400 kit and assembled electronic products — including a new personal LORAN navigational computer suitable for boating or backpacking — are showcased in the latest Heathkit catalog.

Many new products are featured; for example, Heath's instrument line has been expanded to include the ID-4801 EPROM Programmer,

used to program, duplicate, verify, and simulate single-power supply 2500 and 2700-series EPROMS. A new version of the H/Z-100 Desktop Computer is also available, featuring 8MHz operation and equipped with a minimum of 256K bytes of RAM.

Of special interest to readers is an FCC-registered phone patch that employs a new design and special speech transmission circuits that allow the patch to be directly connected to the phone line, thereby eliminating conventional hybrid transformers used on four-wire to two-wire conversions.

For a free copy of Heathkit's new catalog, contact Heath Company, Dep't 150-585, Benton Harbor, Michigan 49022.

Circle #306 on Reader Service Card.

2-position coax switch

The new MFJ-1702 two-position coax switch features one pole, two output positions, and low insertion loss — less than 0.2 dB. Its maximum frequency range os 500 MHz, and it has less than 20 milliohms contact resistance SO-239 connectors.

The MFJ-1702 is designed for high-performance at a reasonable price. It has a VSWR of 1:1.2 and gets better than 60 dB isolation at 300 MHz and better than 50 dB at 450 MHz. The power rating is 2.5 kw PEP, 1 kw CW. Unused terminals are also automatically grounded for protection from static, lightning, and RF.

Hams will find that they can rely on this durable two-position coax swich because MFJ includes a one-year *unconditional* guarantee and an additional 30-day money-back guarantee if the product is purchased directly from MFJ Enterprises.

For more information, contact MFJ Enterprises, P.O. Box 494, Mississippi State, Mississippi 39762.

Circle #307 on Reader Service Card.

plug-in encoder-decoder

Communications Specialists has introduced another new direct plug-in encoder-decoder for three popular radios. Based on the proven TS-32 programmable encoder-decoder, the TS-32JRC plugs directly into the J.R.C. JHM-45S50, Sonar FM-2112/FM-2114, and Repco RSM. No modifications to the radio are necessary.

The TS-32JRC allows individual selection of all 32 standard EIA CTCSS tones on any of the radios' channels. The send and receive tones may be the same or different on each of the 16 channels. The TS-32JRC is available for immediate delivery from factory stock and sells for $62.95. A catalog is available on request.

For further details, contact Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92665-4296.

Circle #308 on Reader Service Card.

PK-64 packet system

The PK-64 form AEA is the first Packet System with both hardware and software optimized for the Commodore-64 computer.

On the hardware side, the PK-64 includes Western Digital 1935 HDLC chip for full-duplex operation. The modem is based on the Exar 2206 and 2211 chip set including a 6-pole post detection filter for improved HF and VHF performance. The PK-64 is designed for small size and light weight, for convenient use. AEA designed the PK-64 to operate from 12 volts DC for maximum flexibility in powering the unit. An AC Adapter is available for those wishing to use a 115 volt AC power source. The PK-64 will work with virtually any voice transceiver.

The hardware is only half the story. No Packet Radio Controller is complete without appropriate communications software. Existing terminal emulation programs used with present Packet Controllers were not designed with Packet Radio Communication in mind and are not optimum. The PK-64 includes its own MBA-TOR™ style communications software which has been optimized for Packet Radio.

The PK-64 software allows Split Screen operation for more efficient Packet Radio communications. This is a valuable feature since it allows receiving and displaying packets while the operator is typing a message or response to be transmitted. There is a built-in word processor style text editor that allows disk or cassette files to be created, edited, or deleted. Commodore 64 text and executable files may be transferred error-free with the PK-64. Ten command and message buffers are available for traffic, bulletins, or often used connect paths, etc. A software clock is included which automatically logs the connect and disconnect times and dates. PK-64 commands are the same style as the TAPR family of commands.

The price is less than $220.

For more information, contact Advanced Electronic Applications, Inc., PO Box C-2160, Lynnwood, Washington 98036.

Circle #302 on Reader Service Card.